RADAR POCKET BOOK

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RADAR POCKET BOOK

A Compendium of useful Data for Engineers and Operators concerned in the Construction, Installation and Use of Radar Equipment

By
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PREFACE

THE aim of this book is to provide, in a reasonably small compass, information on the basic electrical principles and formulae particularly applicable to radar, together with data on the various parts of a radar installation.

Typical circuit diagrams of the several units of a radar equipment are given, and their mode of operation concisely explained.

Although considerations of space prevent this book from being in the nature of a comprehensive textbook, it is hoped that it covers the subject adequately, even if in a somewhat abbreviated form.

Emphasis has been placed on radar working on centimetric rather than metric wavelengths, the latter being largely obsolete.

The subjects dealt with are suitably sectionalized so that the reader can refer to any particular item without necessarily reading the whole book at one time.

Rationalized M.K.S. units (now officially adopted) have been used in this book except where other units are more convenient. Where the nature of a unit is not specified, it is to be understood to be a rationalized M.K.S. unit.

The author desires to express his appreciation of information furnished from time to time by various Companies engaged in the industry, in particular The British Thomson-Houston Co., Ltd., The Sperry Gyroscope Co., Ltd., Kelvin-Hughes Ltd., Cossor Radar Ltd., Decca Radar Ltd., and Marconi International Marine Communication Co., Ltd.

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CHAPTER I

ELECTROMAGNETIC WAVES

ELECTROMAGNETIC RADIATION travels in free space with a constant velocity of $2{\cdot}9979_2 \times 10^8$ metres per second (usually denoted by c).

An electromagnetic wave has components of electric and magnetic force, which are at right angles to one another, and also to the direction of propagation. For this reason electromagnetic waves are called transverse waves. The plane containing the direction of the electric force and the direction of propagation is called the plane of polarization. For a direction of propagation which is substantially horizontal, the wave is said to be vertically or horizontally polarized according to whether the electric field is vertical or horizontal.

Field Strength

At a point at a distance exceeding a few wavelengths from a radiating aerial, the electric and magnetic forces are in phase and are inversely proportional to the distance. Both vary sinusoidally with time.

Waves travel outwards from a radiating aerial in all directions, and the wavefronts (i.e. the surfaces over which the phase is the same at all points at any instant) are spheres, neglecting small variations near the aerial due to its finite size.

Practical problems normally involve consideration of the energy flowing across only a relatively small area at some distance from the source, and the wavefront may therefore be considered as plane over the small area considered, i.e. the wave may be regarded as a plane wave.

Impedance of Free Space

In free space the strengths of the electric and magnetic fields (E and H) bear a definite relation to each other at any given point and time. Their ratio E/H is equal to $\sqrt{\frac{\mu}{\epsilon}}$ and is called the impedance of free space. It has the dimensions of a resistance. (E is measured in volts per metre and H in amperes per metre.)

ϵ and μ are the absolute permittivity and permeability, respectively, of free space, their values in the M.K.S. system being $\mu = 4\pi \times 10^{-7}$ and $\epsilon = 10^7/(4\pi c^2)$, whence the impedance of free space is $4\pi c \times 10^{-7} = 377$ ohms.

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Air is practically equivalent to free space, the velocity of propagation being about one part in 3,000 lower due to the permittivity of air being slightly greater than that of free space. The difference between the magnetic permeabilities of free space and air, and indeed all the usual non-magnetic materials, is negligible.

Energy Flow

For a plane wave, the rate of energy flow across an area of one square metre is EH watts, where E and H are the R.M.S. values of the electric and magnetic fields.

This can also be expressed as H^2Z or E^2/Z, where Z is the impedance of free space. The energy density is inversely proportional to the square of the distance from the source.

Propagation in Practice

As in the case of light, radar waves in free space travel in straight lines, and the curvature of the earth therefore limits the range of detection in practical cases. As with light, atmospheric refraction causes the waves to follow a path curved towards the earth, thus making the effective radio horizon somewhat farther away than the geometrical horizon.

Water vapour has a greater refractive index at radar frequencies than for light waves, and humidity is therefore of some importance, although not sufficient to cause difficulty in practice. For many purposes calculations can be made by assuming the waves travel in straight lines, but that the radius of the earth is 4/3 times its actual value. For propagation over the sea, a ratio of 2/1 is probably more accurate.

Attenuation

Absorption by atmospheric gases is generally negligible at the frequencies used in radar, although at somewhat shorter wavelengths absorption due to molecular resonance occurs.

Rain

Rain can adversely affect the performance of a radar installation in two ways, viz. :

(1) Absorption by water particles.
(2) Scattering of radiation by water particles.

The second effect, by giving somewhat large "spread out" echoes from a storm, may obscure the echo from an object such as a ship or aeroplane situated within the area of the storm. For small objects, this is likely to be more serious than the first effect, but for large objects, e.g. a large ship, absorption may be relatively more serious.

Snow produces similar effects to rain, but to a smaller degree unless the snow is very heavy.

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Fog

Fog produces considerable attenuation if dense enough to reduce optical visibility to tens of yards or yards.

Superrefraction

Under certain conditions, mainly associated with fine warm weather, when the air at a small height is warmer and drier than at the surface, what is effectively a " duct " exists at or near the surface, extending a few tens of feet upwards.

This " duct " can act as a waveguide, tending to guide the rays round the surface of the earth, so extending the effective horizon distance. In extreme cases echoes from large objects such as cliffs, islands, etc., have been detected at distances of several hundreds of miles with radar sets whose normal maximum range may be only a matter of 10 to 20 miles.

In general, this phenomenon is not detrimental, except for possible confusion resulting from a long-distance echo appearing at apparently short range on a trace corresponding to a transmitter pulse later than the one which gave rise to the echo.

Fading

In addition to fading of the type experienced in radio communication, another type, due to variation of reflecting power with changing aspect of the target, may prove troublesome at times. With aircraft such fading may at times be very rapid, the signal varying perhaps 10 to 1 in strength in a matter of less than a second. This can cause trouble with certain types of radar equipment where it is necessary to compare the strengths of two signals separated by a small interval of time.

Another phenomenon is the random wandering of the effective " centre of reflection " over the area of the target. This is known as " glint " and can be troublesome when accurate direction finding is required.

Reflection of Electromagnetic Waves

A plane wave, incident on a plane perfectly conducting surface of infinite extent, is wholly reflected, the angles of incidence and reflection being equal, as in the case of light. If the wave is vertically polarized it is reflected without change of phase, but if horizontally polarized it undergoes a phase change of 180°.*

Reflection also occurs at the surface of a perfect insulator, the angles of incidence and reflection being equal, but the magnitude of the reflected ray is less than that of the incident because some of the energy is refracted into the insulator. The ratio of the amplitude of the reflected ray to that of the incident

* The terms vertical and horizontal polarization can lead to confusion when the direction of propagation of the radiation is not horizontal. To avoid ambiguity it is preferable to describe vertical and horizontal polarization as parallel and perpendicular, respectively, to the plane of incidence (the plane containing the incident and reflected rays).

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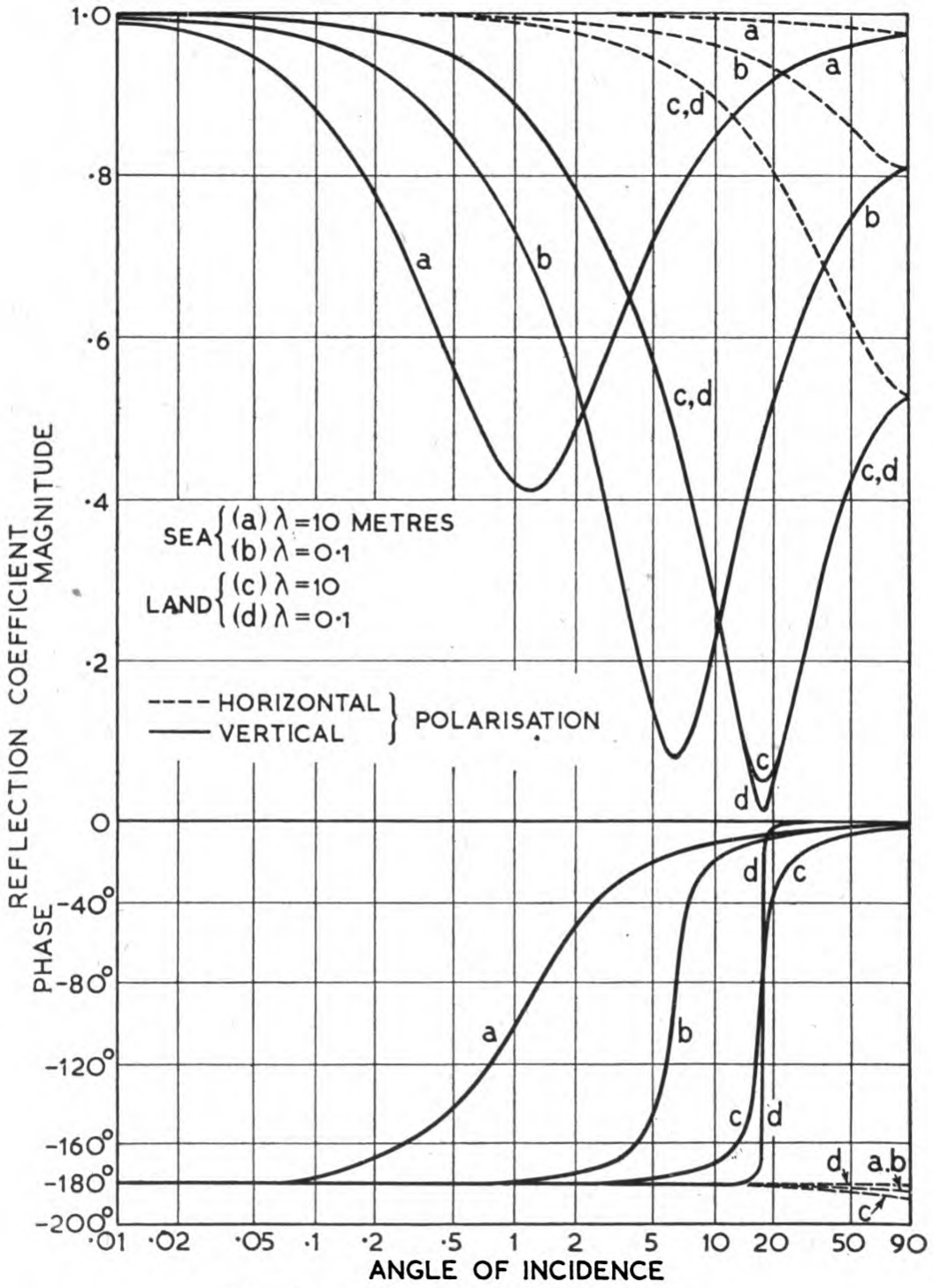

FIG. 1.1.—REFLECTION COEFFICIENT.
Examples for reflection from earth and sea for different wavelengths.

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is given by $\dfrac{\frac{\epsilon_2}{\epsilon_1}\sin\theta - \sqrt{\frac{\epsilon_2}{\epsilon_1} - \cos^2\theta}}{\frac{\epsilon_2}{\epsilon_1}\sin\theta + \sqrt{\frac{\epsilon_2}{\epsilon_1} - \cos^2\theta}}$ for vertical polarization

and $\dfrac{\sin\theta - \sqrt{\frac{\epsilon_2}{\epsilon_1} - \cos^2\theta}}{\sin\theta + \sqrt{\frac{\epsilon_2}{\epsilon_1} - \cos^2\theta}}$ for horizontal polarization,

where ϵ_1 and ϵ_2 are the permittivities of the first and second media, and θ is the angle which the incident ray makes with the reflecting surface.

For the purposes of this section, all materials with which radar is concerned may be regarded as either perfect conductors or perfect insulators, with one important exception, the earth.

Reflection of Electromagnetic Waves from the Earth's Surface

The electrical constants of the earth vary greatly from place to place, and consequently the magnitude and phase of the reflected wave can only be found by calculating each case separately. Different results will be obtained for different wavelengths.

The calculation can be made by substituting for ϵ_2 in the above formulae the complex quantity $[\epsilon_2 - j(\sigma/\omega)]$, where σ is the conductivity of the second medium (i.e. the earth) in mhos per metre, and $\omega = 2\pi f$. ϵ_1 remains unchanged, since it is assumed that the first medium will be free space, or air (which amounts to practically the same thing).

The calculation is laborious, but curves have been published which give the results fairly easily.

To illustrate the way in which reflection varies with different conditions, Fig. 1.1 shows the magnitude and phase of the reflection coefficient for wavelengths of 10 metres and 10 centimetres over sea and "average" land. In calculating these curves, the relative permittivities of land and sea have been taken as 10 and 80, respectively (corresponding to absolute values of $8{\cdot}8 \times 10^{-11}$ and $7{\cdot}1 \times 10^{-10}$), and the specific resistances as 300 and 0·24 ohms per metre for land and sea, respectively. The values for land vary greatly for different types of soil, and the curves in Fig. 1.1 must be taken as illustrative only.

It will be observed that the reflection of horizontally polarized waves is much less affected by the angle of incidence than is the case with vertical polarization.

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CHAPTER II

BASIC RADAR PRINCIPLES

NEGLECTING the effects of atmospheric absorption, reflection from earth or sea, and other disturbing effects, the maximum working range of a radar installation is:

$$r = \sqrt[4]{\frac{W_t G A_e A_r}{W_r 16\pi^2}} = \sqrt[4]{\frac{W_t K A_t A_e A_r}{W_r 16\pi^2 \lambda^2}}$$

where r = range,
W_t = transmitter power,
W_r = minimum input power to receiver to give a workable signal,
A_t = effective area of transmitting aerial,
A_r = ,, ,, ,, receiving aerial,
A_e = ,, echoing area of target,
G = power gain of transmitting aerial,
$K = G\lambda^2/A_t$ (a constant for the particular type of transmitting aerial used),
λ = wavelength.

The above equation shows that

(i) Range is proportional to the fourth root of transmitter power.
(ii) Doubling W_t only increases range by about 20 per cent.
(iii) To double the range, W_t must be increased 16 times.
(iv) Range is increased by decreasing the wavelength. This is because the beam becomes narrower, and G is consequently increased.

Definitions

The effective echoing area of target (A_e) is the area of an hypothetical target, normal to the incident beam, which re-radiates equally in all directions all the energy incident on it, and produces at the receiver a signal equal to that produced by the actual target.

The value of A_e for an actual target depends on many factors, and can only be found empirically. For an average aircraft it is of the order of 1 to 10 square metres.

Transmitter aerial gain (G) is the ratio of power radiated in the direction in which it is a maximum to what it would be if the same total power were radiated equally in all directions.

Effect of Ground Reflections

For targets at low angular elevations, e.g. ships or low-flying aircraft, rays reflected from the earth or sea interfere with the

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direct ray, and the basic equation becomes:

$$r = \sqrt[8]{\frac{W_t K A_t A_e A_r 16\pi^2 h^4 H^4}{W_r \lambda^6}}$$

where h and H are the heights above earth or sea of the radar aerial and target, respectively.

This formula only applies provided r is much greater than $\sqrt{hH}$, and it is based on the assumption of perfect reflection from earth or sea, with 180° change of phase; this assumption is approximately correct for low values of angular elevation.

It should be noted that, in these circumstances, the range is proportional to the eighth root of the transmitter power, which

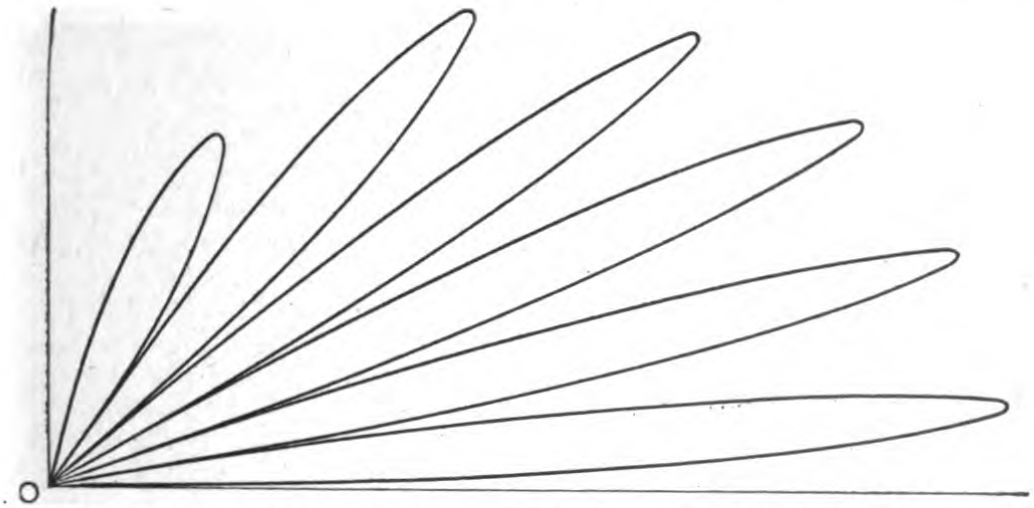

FIG. 2.1.—POLAR DIAGRAM AS MODIFIED BY REFLECTIONS FROM EARTH'S SURFACE.

explains the difficulty of detecting low-flying aircraft at reasonably long ranges.

At greater elevations the problem becomes more complex because, as the elevation increases, the difference between the path lengths of the direct and reflected rays increases, giving a series of maxima and minima as the effective difference becomes an even or odd number of half-wavelengths. The assumption of perfect reflection with phase reversal also becomes less valid.

The general effect is indicated by the polar diagram shown in Fig. 2.1 which has been calculated for an assumed arbitrary case. The field strength in any direction is proportional to the distance of the curve from the point *O*.

With narrow beams and angles of elevation such that ground reflections do not occur, the free space equation is applicable. This will be the case for most centimetric radars at elevations exceeding a few degrees.

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Wavelength

Early radars were designed to work on metre wavelengths, mainly because apparatus and techniques for shorter waves were not available. Most modern radars work on centimetre wavelengths, the 10- and 3-cm. bands being very commonly used.

The basic equation shows that range increases as the wavelength is decreased, because, for a given aerial size, the beam becomes narrower. A narrow beam is desirable for accurate location, e.g. for fire-control purposes, but a wider beam is desirable for searching.

If the required beam width is specified, the equation can be written

$$r = \sqrt[4]{\frac{W_t K \lambda^2 A_e}{256 W_r \theta^4}}$$

where θ is the beam width (in radians) to half-power points. This formula is based on the assumption of a dipole in a circular parabolic reflector, for which $\theta = \lambda/(\text{diameter})$, and that the same aerial is used for transmission and reception.

This form of the equation shows that, for a specified beam width, the range increases as the wavelength is increased. There is a practical limit beyond which increase of wavelength is not possible because, for a given beam width, the diameter of the aerial must be increased proportionately to the wavelength.

In practice, the wavelength must be settled by a suitable compromise, taking account of the various factors involved. Present practice tends towards the use of 10 to 50 centimetres for long-range searching or warning radars where sensitivity at long range is more important than accurate location. For accurate work at shorter ranges, wavelengths of 3 to 10 centimetres are used in most cases.

Pulse Length and Repetition Frequency

The pulse length must be kept short enough to prevent the merging of echoes from two targets which, although close together, are required to be distinguished as separate objects. The echoes from two targets on the same bearing and 150 metres apart will just begin to merge with a pulse length of one microsecond. They would be recognized as two objects by the apparent doubling of the length of the echo on the display, but this would become more difficult as the separation decreased. For accurate work pulse lengths of 0·1 to 1 microsecond are usual, but other considerations apply to long-range sets.

The minimum power required at the receiver depends on noise level, and since noise is present all the time, it is the average power of the transmitter, rather than the peak, that is important for long-range warning sets where the received signal is relatively weak. In such cases pulse lengths up to about 10 microseconds may be used to give greater average power.

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The time separation between successive pulses must be great enough to avoid confusion due to an echo signal, produced by a particular transmitter pulse, appearing on the display trace belonging to the next pulse. Long-range sets must therefore have low pulse repetition frequencies, say about 200 to 500 per second.

For shorter range sets, using short pulse lengths for accurate determination of range, higher pulse repetition frequencies are desirable, and frequently necessary in order to keep up the *average* power. For these installations, figures of 1,000 to 2,000 per second, or more, may be used.

Range Measurement

The basis of range determination by radar is the measurement of the time elapsing between the emission of a pulse from the transmitter and the receipt of the corresponding echo. This time, multiplied by half the velocity of propagation, gives the range.

1	microsecond corresponds to 149·85 metres or 163·88 yards (in air)
6·673	microseconds correspond to 1 kilometre
10·739	,, ,, ,, 1 statute mile
12·366	,, ,, ,, 1 nautical mile (6,080 yards)

Variation of the velocity in air, due to atmospheric changes, is too small to be of practical importance.

Comparatively simple equipments will give range accuracies of about 2 to 3 per cent, but more elaborate circuits, with calibration arrangements, will give results ten or twenty times better than this.

The maximum range obtainable with a radar set depends on the purpose for which it is to be used. Very great ranges have been obtained experimentally (such as echoes from the moon), but it is unusual in practice to require a long-range set to operate at more than 100 to 200 kilometres, or an accurate set (e.g. for fire control) at more than about a fifth of this distance.

Bearing and Elevation

In older radar sets, employing metre wavelengths, bearing and elevation were generally determined by comparing the phase of the echo signal as received on two or more dipoles spaced apart.

With centimetric radars, bearing and elevation are determined by directing the radar beam at the target and noting its direction for maximum strength of echo signal. Accuracies of about 2 to 5 degrees are obtainable in this way. More elaborate methods of determining the direction of the aerial for maximum signal strength enable measurements to be made to a fraction of a degree (see Chapter X).

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CHAPTER III

THERMIONIC TUBES, CRYSTAL RECTIFIERS AND TRANSISTORS

THE electron is a very small particle, having a mass (m) of $9{\cdot}11 \times 10^{-31}$ kilogramme (i.e. about 1/1,840 of that of the hydrogen atom). It carries a negative charge (e) of $1{\cdot}602 \times 10^{-19}$ coulomb. The ratio of charge to mass (e/m) is known more accurately than either e or m separately, and is $1{\cdot}759 \times 10^{11}$ coulombs per kilogramme.

The value given for m applies to low velocities because, by the principle of relativity, the effective mass of a moving particle increases as the velocity increases according to the relation

$$\frac{m'}{m} = \frac{1}{\sqrt{1 - (v/c)^2}}$$

where m is the mass at zero velocity, m' that at velocity v, and c is the velocity of light (approximately 3×10^8 metres per second). The increase is relatively unimportant unless the velocity is an appreciable fraction of that of light; e.g. if $v = c/10$, the increase is only 0·5 per cent.

Electrons can be liberated by heating the emitting body (thermionic emission) or by allowing electrons to strike a body with sufficient velocity to knock out one or more "secondary electrons" (secondary emission).

Thermionic Emission

When a body is heated, the average velocity of the electrons contained in it increases, and any electron possessing sufficient velocity to overcome the electric force tending to hold it back at the surface of the body will escape. The minimum velocity for escape is known as the "work function" and is expressed in a unit known as the "electron volt", which is the velocity acquired by an electron in moving through a potential difference of one volt in an electric field.

Dushman's formula for the emission from a hot body is

$$AT^2 e^{-\alpha \frac{W}{T}} \text{ amperes per square centimetre.}$$

A and α are constants, T is the absolute temperature (in degrees Kelvin) and W is the work function.

For tungsten $A = 60$ $\alpha = 11{,}600$ $W = 4{\cdot}54$ volts.

This formula applies to a valve in which sufficient voltage is applied to the anode to remove all the electrons liberated from the cathode. In practice this condition is not usually fulfilled, and the cloud of electrons (space charge) which collects in the

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neighbourhood of the cathode repels electrons about to leave the cathode, thus reducing the emission.

For this condition, Child and Langmuir have shown that for cathode and anode in the form of parallel plates d centimetres apart the emission current is $2{\cdot}32\frac{V^{\frac{3}{2}}}{d^2}$ microamperes per square centimetre, where V is the voltage between the electrodes, provided the temperature is sufficiently high to avoid any limitation according to the Dushman formula.

Emission from Valve Cathodes

The common cathodes are (*a*) pure tungsten, (*b*) thoriated tungsten (containing a small quantity of thorium oxide) and (*c*) oxide coated (the coating usually consisting of a mixture of barium and strontium, and sometimes calcium, oxides).

Pure tungsten is only used for special purposes, e.g. high-voltage rectifiers, when no other material is suitable. The commonest cathode is the oxide-coated type.

Emissions obtainable with various cathode materials:

Material	Operating Temperature	Emission (amps. per sq. cm.)	Milliamps per Watt
Pure tungsten . .	2,500° K.	0·26	4
Thoriated tungsten .	1,900–2,000	1·2–2·8	50–100
Oxide coated . .	1,000–1,100	0·5–1·6	200–500

The above table gives a rough indication of the peak emission obtainable in the absence of space-charge limitation. Actual figures may differ considerably from those given, depending on constructional details, operating temperature, etc. Oxide-coated cathodes can give from 10 to 20 times the above-quoted figures when operated under pulsed conditions.

Valves for Radar

Modern radar installations operate at frequencies of the order of 1,000 to 10,000 Mc/s, and valve requirements consequently differ somewhat from those for conventional radio communication purposes.

In practice, only a limited number of valves are required to deal with these frequencies. On the transmitting side the oscillating valve is normally the last in the train, subsequent amplification not being employed. On the receiving side, it is now usual to convert to intermediate frequency with a crystal diode, without signal frequency amplification. Some earlier radars employed signal frequency amplification, and this necessitated special designs in order to obtain reasonable efficiency at the high frequencies involved.

The remaining requirements can generally be met by using

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standard radio valves, since the frequencies involved do not exceed about 50 Mc/s (the I.F.), and in many parts of the circuit are considerably lower. It must be noted, however, that currents and voltages in some parts of the circuit are only applied in the form of pulses with relatively long resting times between pulses. Therefore the valves must be capable of standing high peak currents and voltages, although the average loading may be small.

Limitations of Triodes at High Frequencies

At moderately high frequencies the performance of a triode is impaired by the Miller effect; this is the effective increase of input (i.e. grid-cathode) admittance, and reduction of stage gain due to feedback via the anode-grid capacitance.

At still higher frequencies, the inductance of the electrode leads resonates with interelectrode capacitance. A further complication is due to the fact that, owing to electron transit time, changes in the density of the electron stream reaching the anode are no longer in phase with the changes in grid voltage causing them. This results in a deterioration in performance and ultimate failure to function at all when the frequency becomes very high.

Disc Seal Valves

Fig. 3.1 shows diagrammatically the construction of a valve designed to minimize the above effects. The anode and grid leads are in the form of circular discs sealed through the glass envelope. This gives low inductance and also assists in heat

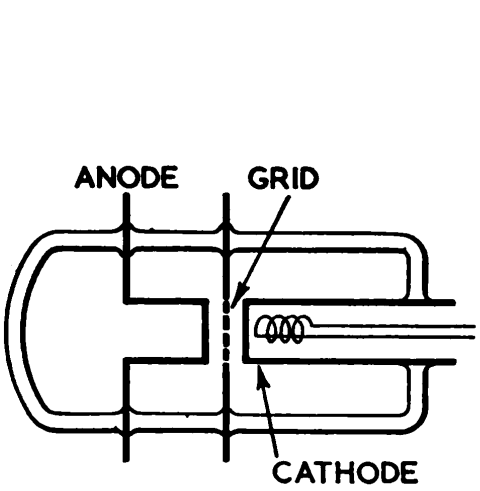

FIG. 3.1.—DISC-SEAL VALVE.

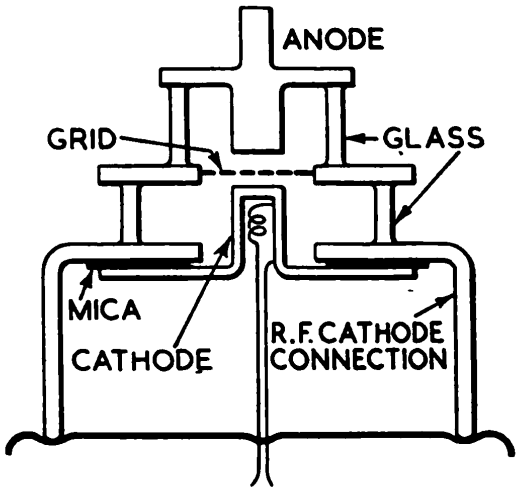

FIG. 3.2.—LIGHTHOUSE VALVE.

dissipation. The layout of the electrode system helps to keep capacitance low, and short electron transit times are obtained by making the interelectrode spacings very small, often less than a millimetre.

Electron transit time can also be reduced by increasing the anode voltage and by working at a high space charge density.

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As most radar valves to which these considerations apply work under pulsed conditions, with relatively long resting times, it is possible to use much higher values of current and voltage than would be permissible with c.w. working.

As the heater power is on continuously, it represents a considerable percentage of the total load, and the use of an oxide-coated cathode, which only requires a relatively small heater power, helps to ease the heat dissipation problem.

Lighthouse Valve

This valve, of American origin, is shown diagrammatically in Fig. 3.2. It is designed to possess advantages similar to those of the disc seal valve.

Performance of Special Triodes

Valves designed on the lines indicated above can give, when operating with suitable push-pull circuits, pulsed outputs up to about 200 kilowatts per pair at 600 Mc/s. Others have been built to give about half this power at double the frequency.

Smaller valves will give pulsed outputs of 500 watts (peak) at 3,000 Mc/s, and will also operate as amplifiers, with reasonable efficiency, at this frequency.

If the input is applied between grid and cathode, and the output taken between grid and anode, the grid will act as a screen between anode and cathode, minimizing feedback from output to input. This is known as the " grounded grid circuit ".

THE KLYSTRON

The klystron is a valve whose action depends on electron transit times, using the principle of velocity modulation. The

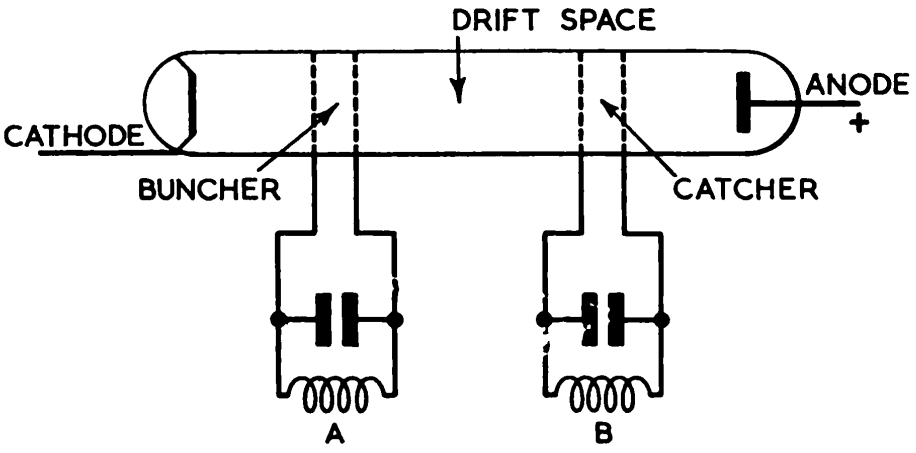

FIG. 3.3.—DIAGRAM ILLUSTRATING PRINCIPLE OF KLYSTRON.

basic principles of operation can be explained by aid of the diagram in Fig. 3.3 which shows a tube having a cathode, an anode, and two pairs of grids, each with its own resonant circuit.

Electrons emitted from the cathode pass through the first pair

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of grids, called the " buncher " grids, where they experience acceleration or retardation, depending on the phase of the oscillation assumed to exist in circuit *A*. Different electrons will therefore leave the buncher grids with different velocities, and will in consequence bunch together into groups as they pass along the drift space.

As the electrons pass the second pair of grids (the " catcher " grids) in bunches, an alternating voltage is induced in the second resonant circuit *B*.

If the device is properly proportioned it will act as an amplifier, the input being applied to *A* and the output taken from *B*. If suitable feedback is provided from *B* to *A*, the arrangement will act as an oscillator.

By using resonant cavities, frequently called " rhumbatrons ", instead of circuits consisting of lumped inductance and capacitance, oscillators capable of giving several kilowatts at frequencies of some thousands of megacycles can be constructed. Owing to the superiority of the magnetron, such oscillators have not, however, been used to any great extent in radar.

The Reflex Klystron

This is shown diagrammatically in Fig. 3.4. The two resonant circuits have been replaced by a single resonant cavity, which now acts also as anode. The original anode has given place to

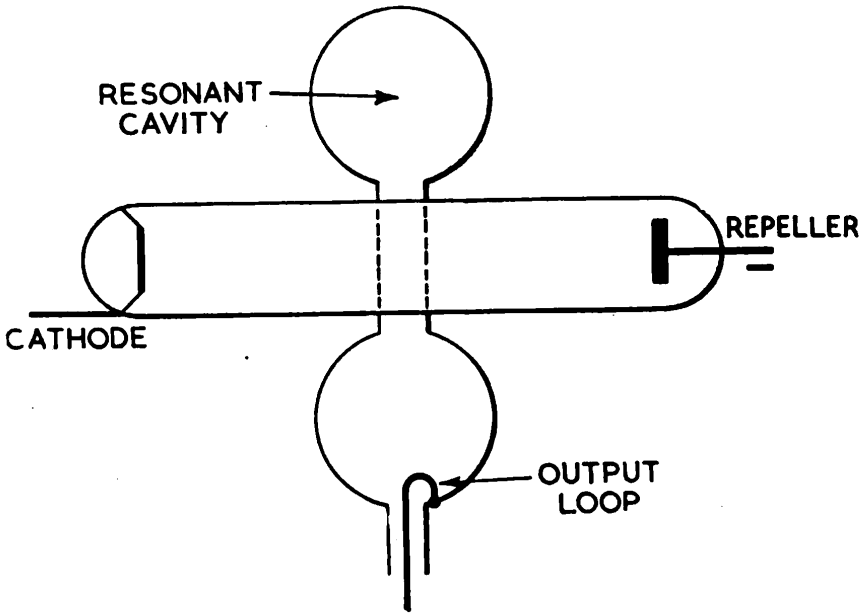

FIG. 3.4.—REFLEX KLYSTRON.

an electrode, called the reflector or repeller, which is maintained at a potential negative to the cathode.

The electrons pass through the grids as in Fig. 3.3, and are then turned back by the repeller and retrace their paths, passing through the grids (now acting as catchers) a second time, and finally being collected on the walls of the resonant cavity.

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As an oscillator, the reflex klystron possesses the advantage of having only one tuned circuit to be adjusted. Also it is possible to make small adjustments to the frequency by altering the voltage on the repeller. Making this more negative increases the frequency. The tuning range obtainable in this way is small, of the order of about one per cent, but is sufficient for small adjustments, and is particularly useful for A.F.C. purposes.

Methods of Construction

For wavelengths in the 10-cm. band, the disc seal type of construction is satisfactory (see Fig. 3.5). The part of the resonator inside the envelope is made re-entrant, the apertures acting as grids (actual grids not normally being used). A large part of the resonator is outside the envelope and is completed by a brass cylinder between the two discs forming the top and bottom of the cavity. From four to six brass plugs are screwed into the cylinder to provide coarse tuning. Power is taken out by means of a coupling loop inside the cavity.

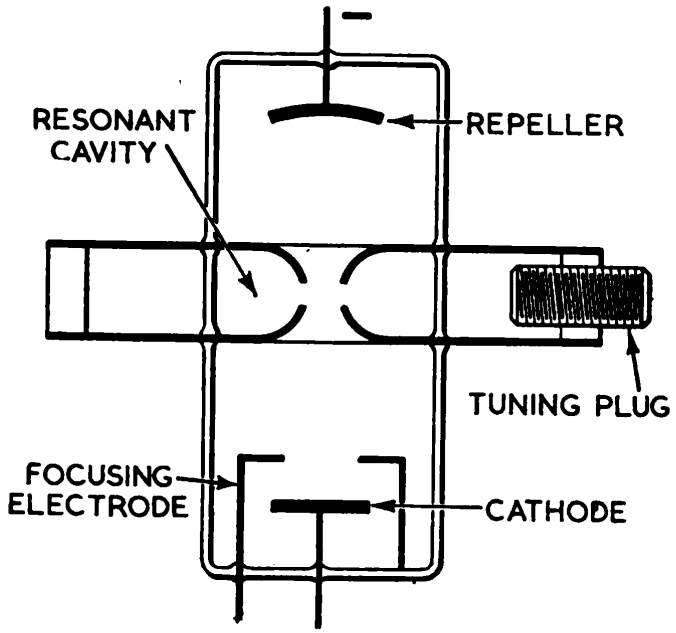

FIG. 3.5.—REFLEX KLYSTRON FOR 10-CM. BAND.

For wavelengths in the 3-cm. band the whole of the resonator is inside the glass envelope and coarse tuning is effected by altering the capacitance between the two apertures by moving them relatively to one another; the central part of the cavity, containing the apertures, is made sufficiently flexible to permit this. A flexible diaphragm forming the top of the envelope allows the necessary movement to be imparted from outside.

Power Input and Output

These vary with different types of valve, and the following figures are only intended to give a rough idea. For a 10-cm. klystron, an output of 100 to 200 milliwatts can be obtained with an input of about 5 to 10 watts at 250 volts, the repeller being maintained about 150 to 200 volts negative to the cathode. Three-centimetre valves give about a quarter of the above power, but this is quite sufficient for a local oscillator.

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Electronic Tuning and Voltage Modes of the Klystron

A change of one volt on the repeller electrode will change the frequency by about one or two megacycles at 10 and 3 centimetres respectively, the total practicable variation being some 20 to 40 Mc/s.

For maximum power output the transit time of an electron between its first and second passage past the grid or resonator aperture must be approximately $N + \frac{3}{4}$ cycles, where $N = 0$, 1, 2, 3, etc. Different values of N correspond to different "voltage modes". The higher the (negative) voltage on the repeller,

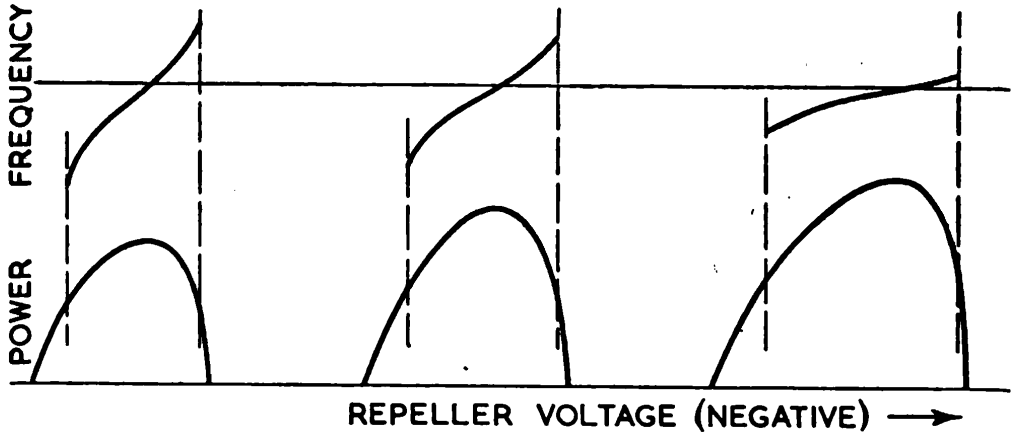

FIG. 3.6.—KLYSTRON MODES.

Showing how power output and frequency vary with change of repeller voltage.

the lower the value of N, and in general the greater the power output. This is illustrated in Fig. 3.6.

The electronic tuning range for any particular voltage mode is defined as the total frequency difference between the two points on the curve (on each side of the maximum) at which the power output is half the maximum for that mode. The lower modes, having greater power, have smaller tuning ranges, and the designer may sometimes have to choose a mode which gives the best compromise.

These voltage modes must not be confused with the different modes of oscillation of a cavity resonator corresponding to different configurations of magnetic and electric fields within the cavity.

THE MAGNETRON

The high-power resonant cavity magnetron is now almost universally used as the oscillation generator in centimetric radar transmitters. It is a transit time oscillator depending for its working on the behaviour of electrons in crossed electric and magnetic fields.

Other types of oscillator, to which the term magnetron has been applied, have not been used to any extent in radar, and

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are not dealt with here. Their mode of operation is roughly similar, but with certain important differences.

Movement of Electron in Electric Field

An electron situated in an electric field of strength V volts per metre is acted on by a force of Ve newtons, where e is the charge on the electron. The direction of the force is parallel to that of the field, and its sense is such as to make the electron move from the more negative to the more positive part of the field. This force produces an acceleration of Ve/m, where m is the mass of the electron. The electron thus tends to move in the same way as a body under the influence of gravity.

Movement of Electron in Magnetic Field

An electron moving with velocity v in, and at right angles to the direction of, a magnetic field of flux density B is subject to a force of magnitude Bev in a direction at right angles to both the magnetic field and the direction of motion of the electron.

Since the force is always at right angles to the direction of motion, a magnetic field, unlike an electric, cannot alter the magnitude of the velocity of an electron, but can only change its direction. An electron in a magnetic field, without any electric field, will, if moving at all, move in a circle whose radius is easily found, by the application of elementary mechanical principles, to be $(mv)/(Be)$. The time taken to travel once round this circular orbit is $2\pi m/Be$, which is independent of the radius and depends only on the strength of the magnetic field. The plane of the orbit is perpendicular to the direction of the field. For field strengths used in actual magnetrons this time is from 10^{-4} to 10^{-3} microseconds. Note that if the electron has an initial component of velocity in the direction of the magnetic field, this component will be unaffected, and when combined with the circular motion will result in the electron traversing a helical path.

Movement of Electron in Crossed Electric and Magnetic Fields

The motion of an electron in electric and magnetic fields which are at right angles to each other is more complex. The electric field tends to make the electron move with ever-increasing velocity parallel to the field, whilst the magnetic field tries to make it move in a circle. The result is a cycloidal path somewhat like that shown in Fig. 3.7, the exact form depending on the initial velocity of the electron. The fundamental point about this curve is that the *average* velocity of the electron is equal to E/B and is horizontal, i.e. at

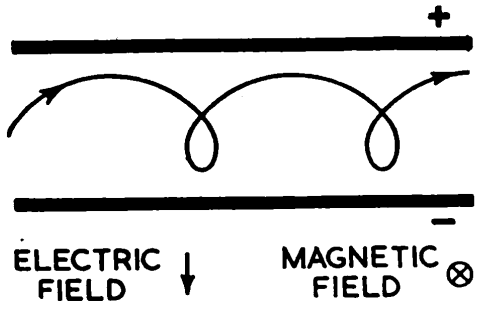

FIG. 3.7.—MOTION OF ELECTRON IN CROSSED MAGNETIC AND ELECTRIC FIELDS.

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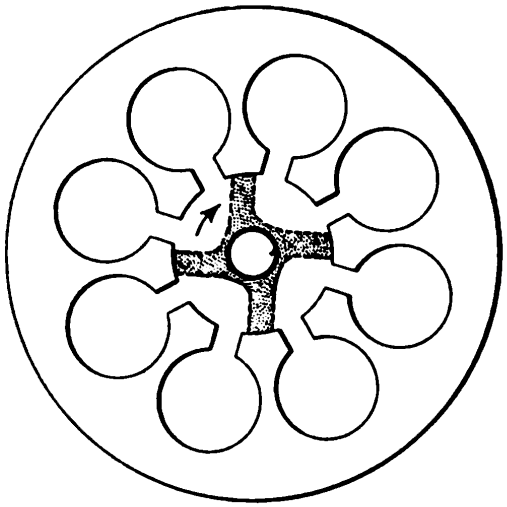

FIG. 3.8.—MAGNETRON.
Diagrammatic representation of cavities and electron cloud.

right angles to both fields. The vertical component is, on the average, zero. The electron is not, as might be expected, caused to travel, *on the average*, towards the positive electrode producing the electric field; on the contrary, it tends to travel along an electric equipotential.

Magnetron Construction

The electrode system of a resonan cavity magnetron is shown diagrammatically in Fig. 3.8, which also shows the electron cloud, to be referred to later.

The anode and cathode are cylindrical, the former being made of high-conductivity copper with a number of slots connected to resonant cavities as shown. These cavities, having distributed capacitance and inductance, form the tuned circuit. A typical method of construction is shown in Fig. 3.9. An axial magnetic field is provided by an external permanent magnet. This is sometimes built in as an integral part of the magnetron, which is then called a " package magnetron ".

Maintenance of the correct field strength within fairly narrow limits is necessary, and the magnet must therefore be carefully handled in practice; it must not be knocked, nor allowed to come into contact with magnetic materials, e.g. tools.

Power take-off is usually by means of a coupling loop in one of the cavities, although in some cases, particularly at the shorter wavelengths, a slot in one of the cavities provides direct coupling to a waveguide. As all the cavities are fairly tightly coupled magnetically through the end spaces, it is only necessary to couple the output to one cavity.

Magnetron Cathodes

Very high peak emissions are required, and for this purpose the oxide-coated cathode is very suitable as it can give high emission under pulsed conditions. Another important factor is secondary emission produced by those electrons which, after leaving the cathode, return to it with appreciable velocity and liberate further electrons.

The bombardment of the cathode by returning electrons produces appreciable heating, and with some magnetrons it is necessary to reduce the heater current, once the magnetron has started to work, in order to prevent damage to the cathode by overheating.

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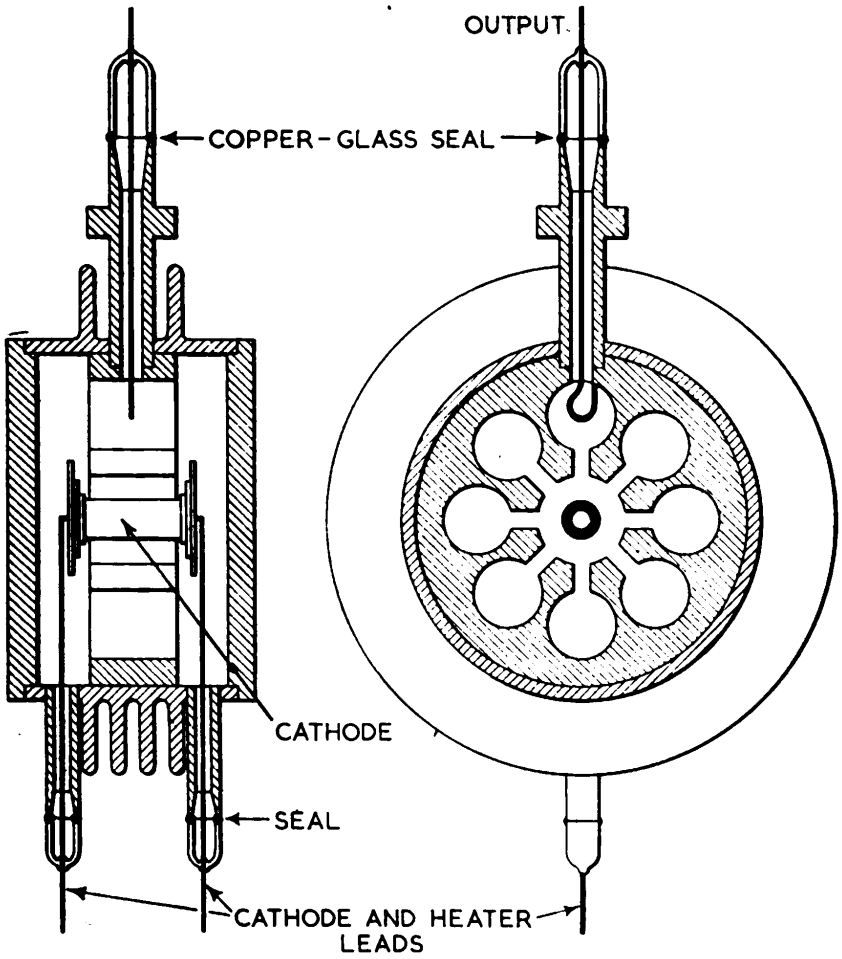

FIG. 3.9.—TYPICAL MAGNETRON CONSTRUCTION.

Principle of Magnetron Operation

In an oscillating magnetron there is a steady d.c. electric field applied between anode and cathode, and an R.F. field due to the oscillation between adjacent anode segments. At any instant successive segments are, from the R.F. point of view, alternately positive and negative.

It can be shown that the space charge cloud takes the form of a spoked wheel which rotates at such a speed that each spoke passes two anode segments per cycle of oscillation—see Fig. 3.8. Of the individual electrons in this cloud, some return to the cathode soon after leaving it (due to the magnetic field causing their paths to curve back towards the cathode), whilst others, due to having somewhat greater velocities or being more favourably placed, ultimately reach the anode.

Of the latter, some arrive at segments which are, at the time, at a positive R.F. potential ; others arrive at negative segments. The former have work done on them by the R.F. field, but the latter are opposed by this field, and consequently give up some

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of their energy to it. It can be shown that the number of electrons arriving at negative (R.F.) segments is greater than the number arriving at positive segments. From this it follows that the R.F. field experiences a net gain of energy, derived initially from the action of the d.c. field on the electrons. Oscillations can therefore be maintained, at a frequency mainly determined by the resonant frequency of the cavities.

Cut-off, Threshold, and Instability Voltages

In the absence of oscillation, the tendency for an electron path to be curved towards the cathode by the action of the magnetic field prevents any electrons reaching the anode unless the anode-cathode voltage exceeds a certain value, called the Hull cut-off voltage. For a cylindrical magnetron the cut-off voltage is

$$\frac{eB^2r_a^2}{8m}\left[1 - (r_c/r_a)^2\right]^2$$

where r_a and r_c are the radii of the anode and cathode.

With a resonant cavity magnetron, however, any slight disturbance can cause oscillation to start provided the voltage exceeds the " instability voltage ". Once started, the oscillation can be maintained if the voltage exceeds the " threshold voltage ". Both these voltages are less than the cut-off voltage.

Threshold voltage is given by

$$\frac{\pi c r_a^2}{n\lambda}\left[1 - (r_c/r_a)^2\right]B - \frac{2\pi^2c^2r_a^2m}{e\lambda^2n^2}$$

where c = velocity of light, λ = wavelength, n = mode number. Instability voltage is roughly, but generally not exactly, equal to threshold.

Modes of Oscillation

Normally the R.F. voltages on adjacent anode segments differ in phase by π radians, i.e. one half-cycle. The number of complete waves of R.F. potential round the anode face is therefore $n = N/2$, where N is the number of segments. This mode of oscillation is called the π mode.

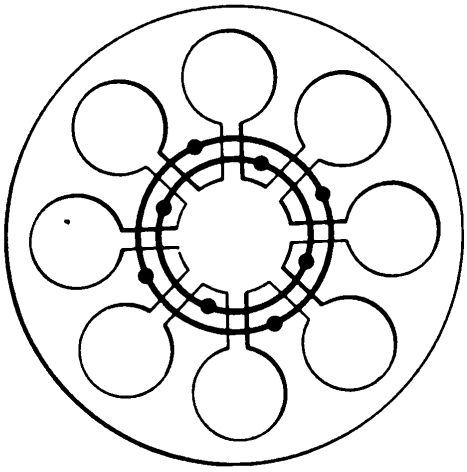

FIG. 3.10.—MAGNETRON STRAPPING.

A system consisting of N separate resonant systems coupled together can, however, oscillate in N ways, each of which corresponds to a different potential distribution round the anode, the phase difference between adjacent segments having values other than π. These other modes are objectionable, largely because each may occur on either of two nearly but not quite equal

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frequencies. This leads to possible jumping from one mode to another, with consequent inconsistency in operation. The π mode corresponds to a single frequency, and is always used in practice. The frequencies of all modes are roughly equal.

By connecting alternate segments together, as in Fig. 3.10, operation in the π mode is made more certain. This is called strapping.

Effect of Load—Frequency Pulling Figure

If the load and feeder are not properly matched to the magnetron the power and frequency are affected to an extent depending on the magnitude (standing wave ratio) and phase of the mismatch.

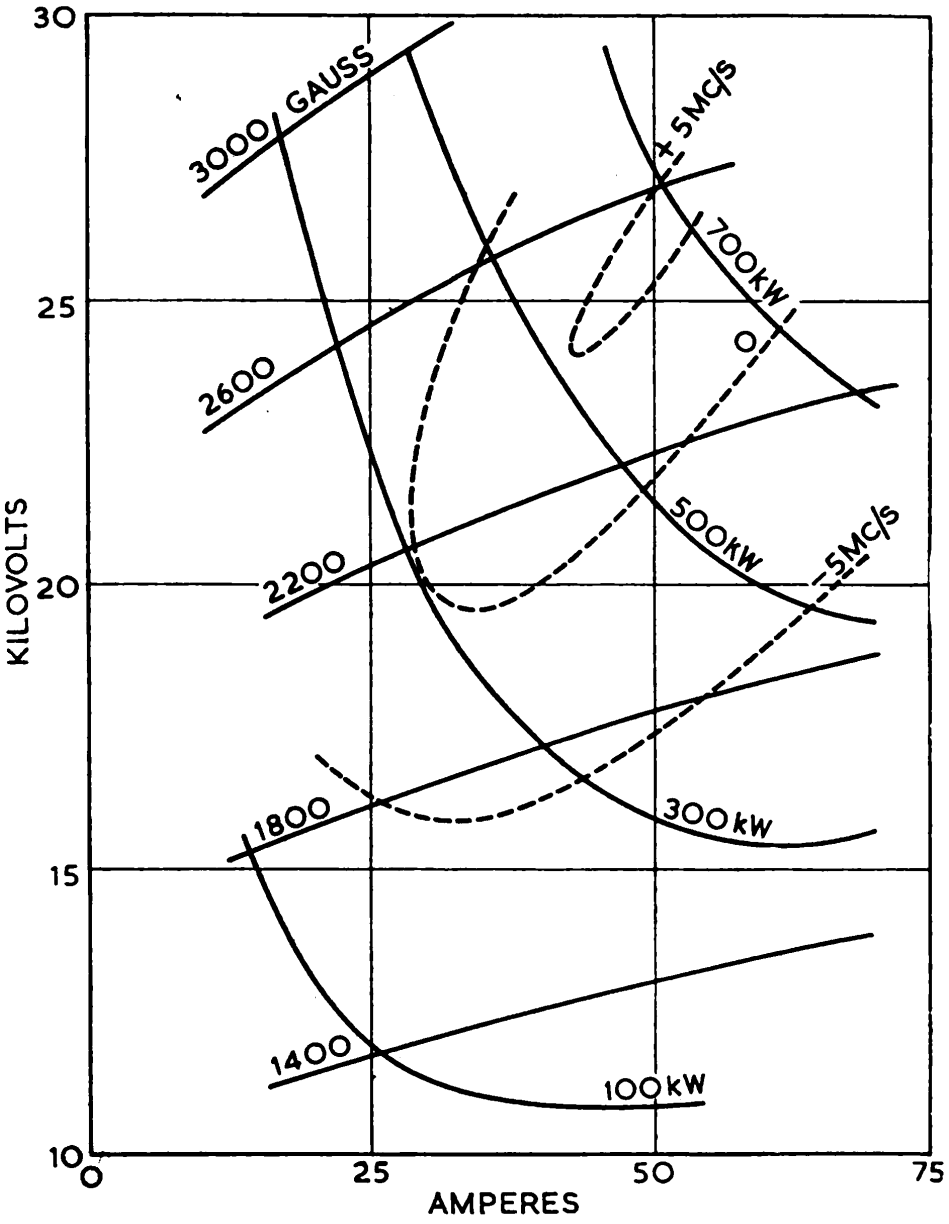

FIG. 3.11.—TYPICAL PERFORMANCE CHART OF MAGNETRON.

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In practice, with a given feeder, there will always be some uncertainty as to the exact value of the mismatch, but it is customary to assume that, in a particular installation, the standing wave ratio will never be worse than a certain figure, but that it may have any phase. The total variation of frequency which may be expected due to load variations within an upper limit of S.W.R. of 1·5 is called the " frequency pulling figure ". In practical cases this may be about 0·1 to 0·2 per cent.

Input Power

The effects of variation of input power, magnetic field, etc., are illustrated in Fig. 3.11. This diagram, called a " performance chart ", is for a particular magnetron. The charts for different magnetrons naturally differ, but all are of the same general nature. The frequency varies with power; the variation in megacycles per second per ampere change of input is called the " frequency pushing figure ".

Frequency Spectrum of Magnetron

A magnetron working under pulsed conditions produces sidebands as well as carrier, as in the case of any modulated oscillator. Theoretically these sidebands extend from zero to infinity, but

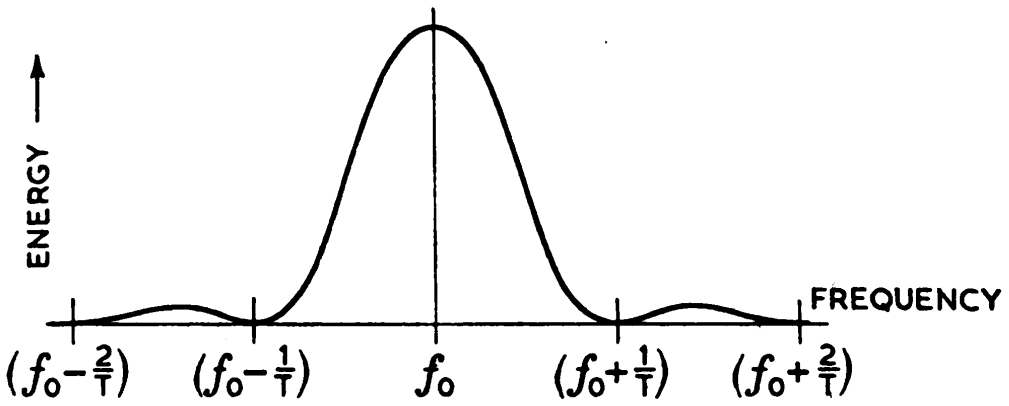

FIG. 3.12.—FREQUENCY SPECTRUM OF PULSED MAGNETRON.

most of the energy is concentrated in a band of width $2/T$ megacycles, where T is the duration of the pulse in microseconds. This is illustrated in Fig. 3.12. The receiver must have a bandwidth of at least $2/T$ if the pulse is to be reproduced with reasonable fidelity. See "Single (Non-recurrent) Pulse"—page 31.

GASFILLED TRIODES (THYRATRONS)

Thyratrons are used in radar mainly as " switches " in the modulator circuit, although lower power versions are used in some ancillary control circuits.

If the grid is held sufficiently negative, no current will flow; as the grid voltage is reduced, a point is reached at which anode

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current begins to flow, as in an ordinary triode. Ionization of the gas then occurs, and the current builds up to a large value (arc discharge), and the anode voltage, i.e. the voltage drop across the arc, falls to a comparatively low value.

The ratio of anode voltage to negative grid voltage at which the discharge starts is constant over a large part of the characteristic, and is known as the control ratio.

Once the discharge has started, the grid loses control, and the arc can only be quenched by reducing the anode voltage below the ionizing potential for the particular gas or vapour employed. An exception to this general law is that an arc of low current density, with a low gas pressure, can be quenched by the application of a sufficiently large negative potential to the grid.

The commonest gases used are mercury vapour and hydrogen. With the former the arc current cannot rise at a greater rate than about 1,000 amperes per microsecond, and the deionization time limits the possible pulse repetition frequency to about 1,000 per second.

Control of vapour pressure, and therefore temperature, is important ; too low a pressure results in an unduly long build-up time of the arc, whilst too high pressure reduces the maximum anode voltage which can be " held off ". The usual permissible maximum temperature is about 45° to 60° C.

The hydrogen-filled thyratron, owing to the lower mass of the hydrogen molecule and its correspondingly greater mobility, can deal with rates of rise up to about 2,500 amperes per microsecond, and pulse repetition frequencies of 3,000 to 4,000 per second.

During conduction, the drop across the arc is of the order of 12 to 15 volts for mercury vapour and 60 to 70 for hydrogen.

Under pulsed conditions thyratrons can be made to pass peak currents of several hundred amperes. To limit grid current an additional screening electrode is sometimes inserted between the cathode and grid.

CRYSTAL DIODES

Crystals are generally used as mixers in centimetric radar receivers because their performance is superior to electronic valves at these frequencies.

The best combination for this purpose is a silicon crystal with tungsten "cat's-whisker" built up into a unit as shown in Fig. 3.13. The cat's-whisker is cemented in place and the assembly sealed up. The complete unit is relatively robust and will withstand reasonable handling.

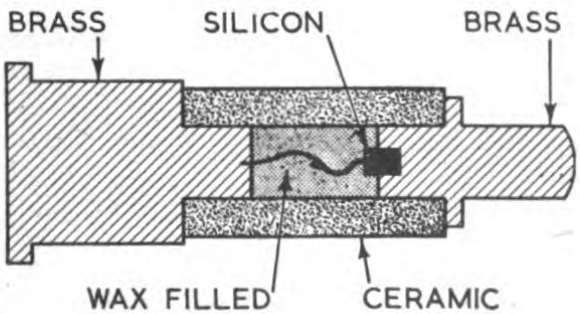

FIG. 3.13.—CRYSTAL DIODE.

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The performance of the crystal is considerably affected by small traces of impurities, controlled quantities of which are introduced in manufacture to obtain certain desired properties. Conversion loss is reduced by the addition of about 0·005 per cent of boron, and the overload capacity can be increased by adding 0·1 to 0·2 per cent of aluminium or beryllium.

The voltage/current characteristic of a typical crystal is shown in Fig. 3.14. Production crystals can readily be made with a backward to forward resistance ratio of about 10, and a forward

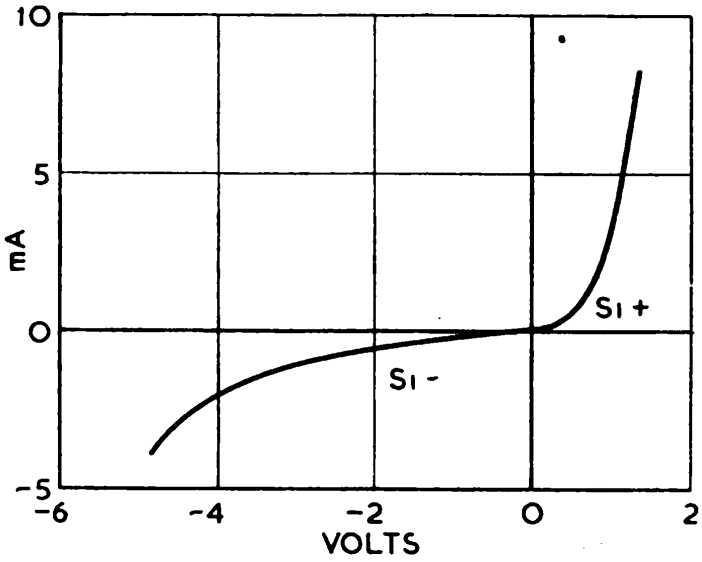

FIG. 3.14.—SILICON CRYSTAL VOLTAGE/CURRENT CHARACTERISTIC.

resistance of about 200 to 300 ohms. These crystals are suitable for use up to at least 10,000 Mc/s.

At low input voltages the crystal behaves as a square law rectifier. When used as a mixer, however, the input from the local oscillator is made large enough to cause the crystal to operate over a small portion of the "straight" part of its characteristic, and so behave as a substantially linear rectifier, with reasonable efficiency. The standing current under these conditions is usually about 0·5 to 1 milliampere.

Overload Capacity of Crystal

A crystal can easily be burned out by excessive current. As the thermal capacity of the contact is very small, the limit is set by the total energy input rather than by the actual current, and in practice one erg (10^{-7} joule) is about the safe limit of power which the crystal can absorb at one time.

Germanium Crystal Diodes

These cannot be made with so small a contact capacitance as the silicon type, and their use is consequently limited to frequencies up to about 100 Mc/s. Below this frequency germanium has a somewhat higher efficiency than silicon, due

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mainly to its lower resistance, and it can stand much higher reverse voltages; up to 100 to 200 volts, compared with about 4 for silicon. The germanium crystal has a useful application as a second detector in a superheterodyne receiver.

Crystal Triodes (Transistors)

In addition to their use as rectifiers, germanium crystals can also be made to act as amplifiers.

Germanium of a very high degree of purity (impurity not more than about 1 part in 10^8) is deliberately adulterated with minute quantities of suitable impurities. The added substances are of various kinds, some of which result in an excess of electrons in the germanium, whilst others cause a deficiency (called "positive holes"). The two resulting types of crystal are called *n*-type and *p*-type germanium, respectively.

One type of transistor, shown diagrammatically in Fig. 3.15 (*a*), has two cat's-whiskers very close together. The input cat's-whisker is called the emitter, and the output one the collector.

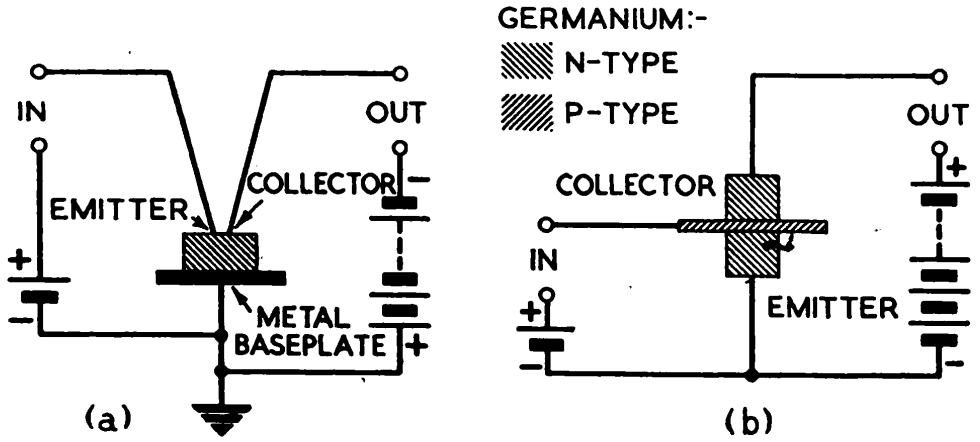

FIG. 3.15.—TRANSISTORS.
(*a*) Point contact type.
(*b*) Sandwich type.

Another form is the sandwich type [Fig. 3.15 (*b*)], consisting of a thin slice of *p*-type germanium between two crystals of *n*-type (or vice versa, with the battery polarities reversed).

The action of transistors is not, as yet, fully understood, but they show promise of superseding thermionic tubes for some applications. The point contact type [3.15 (*a*)] will operate satisfactorily up to 200 or 300 Mc/s, with a power gain of about 20 decibels. The sandwich type [3.15 (*b*)] is not so efficient at high frequencies, but is more effective and stable than the point contact type at medium frequencies, when it will give power gains of about 40 decibels.

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CHAPTER IV

CIRCUIT CALCULATIONS

ALTHOUGH all currents and voltages in a radar installation vary in a periodic manner, in the sense that the pattern of the variations is repeated periodically, it is convenient in practice to use this conception in calculating only some of the circuits involved ; variations in others are more easily dealt with by regarding them as transient effects which take place in such a short time compared with the period elapsing before they are repeated that they may be looked upon as isolated events.

Sinusoidal Alternating Voltages and Currents

The simplest form of a.c. voltage is one that varies sinusoidally with time, i.e. its value at any instant t can be represented by $v = V \sin \omega t$, where v is the instantaneous value, V the peak value, and $\omega = 2\pi f$, where f is the frequency. This equation implies that time is measured from the instant at which the actual voltage is passing through zero. More generally the expression may be written $v = V \sin (\omega t + \theta)$, where θ is an arbitrary phase angle depending on the moment chosen for zero time.

Graphical Representation

In Fig. 4.1 the circle has a radius equal to V, the angle $BOA = \theta$ and $COB = \omega t$. The radius OC may therefore be

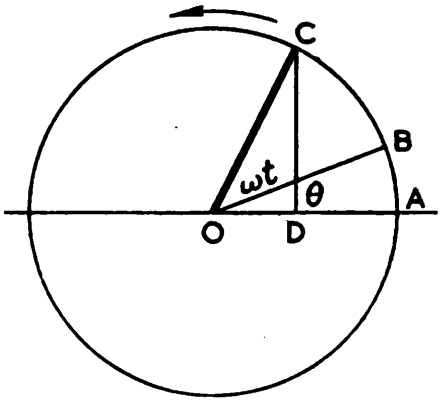

FIG. 4.1.—REPRESENTATION OF SINUSOIDALLY VARYING VOLTAGE OR CURRENT BY ROTATING VECTOR.

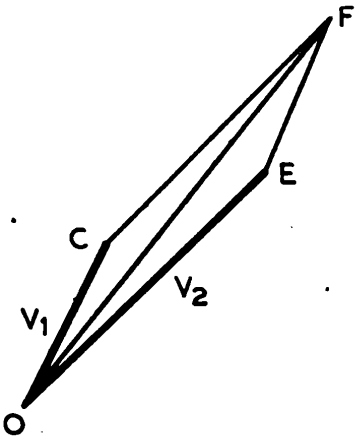

FIG. 4.2.—VECTOR ADDITION OF VOLTAGES.

regarded as continuously rotating in an anti-clockwise direction with angular velocity ω. The instantaneous voltage $v = V \sin (\omega t + \theta)$ is represented by the length of the vertical

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line CD, i.e. the instantaneous value of the voltage is represented by the height of the point C above the centre line at the instant in question.

If there are two different voltages (of the same frequency) but of different magnitudes and phase, say $v_1 = V_1 \sin(\omega t + \theta_1)$ and $v_2 = V_2 \sin(\omega t + \theta_2)$, the sum of these can be found graphically as shown in Fig. 4.2, in which OC, OE and OF represent V_1, V_2 and the resultant, respectively; $COEF$ is a parallelogram, and OC and OE are inclined to the horizontal at θ_1 and θ_2, respectively.

The mathematical solution of this problem is

$$V_1 \sin(\omega t + \theta_1) + V_2 \sin(\omega t + \theta_2) = V_3 \sin(\omega t + \theta_3)$$

where $$V_3^2 = V_1^2 + V_2^2 + 2V_1V_2 \cos(\theta_1 - \theta_2)$$

and $$\tan \theta_3 = \frac{V_1 \sin \theta_1 + V_2 \sin \theta_2}{V_1 \cos \theta_1 + V_2 \cos \theta_2}.$$

Similar methods and considerations apply to currents.

Resistance, Reactance and Impedance

The voltage across a pure resistance R carrying a current $i = I \sin(\omega t + \theta)$ is, by Ohm's law, $v = V \sin(\omega t + \theta)$, where $V = RI$.

In the case of an inductance, L, without resistance, the voltage is proportional to the rate of change of the current, and is $v = \omega LI \cos(\omega t + \theta) = \omega LI \sin(\omega t + \theta + 90°)$. There is a phase change of 90°. The quantity ωL (which depends on the frequency) is called the reactance of the circuit.

For a capacitor C, the current is proportional to the rate of change of voltage, and the expression becomes

$$v = I/(\omega C) \sin(\omega t + \theta - 90°).$$

The quantity $1/\omega C$ is the reactance, and the phase change is negative in the case of a capacitor.

Summarizing, a pure resistance obeys Ohm's law, but when this is applied to inductance or capacitance, the term resistance must be replaced by reactance [ωL for inductance, and $1/(\omega C)$ for capacitance] and a phase change of 90° must be introduced. For an inductance, the voltage leads the current by 90°, and lags it by the same amount for a capacitance. Making due allowance for these facts, the graphical method may be used for calculating the voltages and currents in a circuit containing a mixture of resistance inductance and capacitance.

The Operator j

$j = \sqrt{-1}$. It is called an imaginary quantity. Expressions containing j as a factor are called imaginary, those without j are real, and those in which only a part involves j are called complex.

This mathematical artifice is useful for calculations if we assume that

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$$j \sin(\omega t + \theta) = \cos(\omega t + \theta)$$

and $$j \cos(\omega t + \theta) = -\sin(\omega t + \theta),$$

and take the reactance of an inductor as $j\omega L$ and that of a capacitor as $-j/(\omega C)$. For example, the voltage across a circuit consisting of R, L and C in series, and carrying a current $i = I \sin(\omega t + \theta)$ is

$$[R + j\omega L - j/(\omega C)]I \sin(\omega t + \theta)$$
$$= RI \sin(\omega t + \theta) + I[\omega L - 1/(\omega C)] \cos(\omega t + \theta).$$

This method may be used for calculating the voltages and currents in any circuit in a manner similar to that used for d.c. circuits provided resistance R is replaced, for each circuit element, by $R + j\omega L - j/(\omega C)$, and j is eliminated at the end of the calculation by changing sine to cosine, or vice versa, as required. (*Note:* If the answer contains j^2, this is replaced by -1, similarly $j^3 = -j$, $j^4 = 1$, etc.)

The quantity $R + j\omega L - j/(\omega C)$ is called the impedance of the circuit, and is usually denoted by Z.

$\omega L - 1/(\omega C)$ is the reactance, commonly denoted by X, thus $Z = R + jX$.

Admittance, Conductance and Susceptance

The reciprocal of impedance is called admittance. This is usually denoted by Y and may be written $Y = G + jB$, where G is the conductance and B the susceptance.

$$Y = \frac{1}{Z} = \frac{1}{R + jX} = \frac{R - jX}{(R + jX)(R - jX)} = \frac{R - jX}{R^2 + X^2}$$

whence $$G = \frac{R}{R^2 + X^2} \qquad B = \frac{-X}{R^2 + X^2}$$

conversely $$R = \frac{G}{G^2 + B^2} \qquad X = \frac{-B}{G^2 + B^2}$$

The Exponential $e^{j\omega t}$

It can be shown mathematically that

$$e^{j\omega t} = \cos \omega t + j \sin \omega t$$

where e (= 2·71828 . . .) is the base of the natural or Napierian logarithms.

In a calculation starting with an expression containing cos ωt it is sometimes simpler to use $e^{j\omega t}$ instead, and replace it by its equivalent at the end of the calculation, and then disregard all imaginary terms. Alternatively, we may start with sin ωt, in which case the procedure is the same, but all real terms are disregarded at the end.

Examples

(*a*) Find the voltage across a circuit consisting of R and L in series, carrying a current of peak value I. Let $i = I \cos \omega t$.

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Put $i = Ie^{j\omega t}$, then the voltage is

$$(R + j\omega L)Ie^{j\omega t} = (R + j\omega L)I(\cos \omega t + j \sin \omega t)$$
$$= RI \cos \omega t - \omega LI \sin \omega t + j(\omega LI \cos \omega t + RI \sin \omega t)$$

which, after discarding the imaginary part, becomes simply $RI \cos \omega t - \omega LI \sin \omega t$.

(*b*) Assume a voltage $V \sin \omega t$ is applied across a circuit consisting of R and C in series. Find the current.

In this case, the voltage is represented by the imaginary part of $Ve^{j\omega t}$. The current is

$$\frac{Ve^{j\omega t}}{R - j/(\omega C)} = \frac{V\omega Ce^{j\omega t}}{R\omega C - j}$$
$$= \frac{V\omega Ce^{j\omega t}(R\omega C + j)}{R^2\omega^2C^2 + 1}$$
$$= \frac{V\omega C(\cos \omega t + j \sin \omega t)(R\omega C + j)}{R^2\omega^2C^2 + 1}$$
$$= V\omega C\frac{R\omega C \cos \omega t - \sin \omega t}{R^2\omega^2C^2 + 1} + jV\omega C\frac{R\omega C \sin \omega t + \cos \omega t}{R^2\omega^2C^2 + 1}.$$

Taking only the imaginary part, we have

$$\text{current} = V\frac{R\omega^2C^2 \sin \omega t + \omega C \cos \omega t}{R^2\omega^2C^2 + 1}.$$

The advantage gained is not very apparent in simple examples of this kind, but the saving of labour in a lengthy circuit calculation may be considerable.

Fourier Analysis

An a.c. waveform that is not a simple sinusoid can be represented by

$$V = A_0 + A_1 \sin \omega t + A_2 \sin 2\omega t + A_3 \sin 3\omega t + A_4 \sin 4\omega t + \ldots$$
$$+ B_1 \cos \omega t + B_2 \cos 2\omega t + B_3 \cos 3\omega t + B_4 \cos 4\omega t + \ldots$$

The term A_0 represents a d.c. component, and is zero when the areas of the positive and negative halves of the wave are equal. The term or terms containing $\sin \omega t$ and/or $\cos \omega t$ are called the fundamental, whilst the remainder are the second, third, fourth, etc., harmonics. The frequencies of the harmonics are exact multiples of that of the fundamental. If the shape of the wave is known, the values of the coefficients A and B can be calculated as follows:

$$A_0 = \frac{\omega}{2\pi}\int_0^{2\pi/\omega} v\,dt \qquad A_n = \frac{\omega}{\pi}\int_0^{2\pi/\omega} v \sin n\omega t\,.dt$$

$$B_n = \frac{\omega}{\pi}\int_0^{2\pi/\omega} v \cos n\omega t\,.dt$$

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these calculations being carried out for all values of n from one upwards. In the most general case, there will be an infinite number of terms, but in practice it is usually sufficient to take only the first few, the number actually taken depending on the particular case.

Analysis of Square and Sawtooth Waves

The simple square wave shown by the full line in Fig. 4.3 is represented by the Fourier series:

$$v = \frac{4V}{\pi}\left(\sin\theta + \frac{\sin 3\theta}{3} + \frac{\sin 5\theta}{5} + \frac{\sin 7\theta}{7} + \ldots\right)$$

the summation being made for every odd value from one to infinity.

Analysis of the sawtooth wave shown in Fig. 4.4 gives

$$v = \frac{2V}{\alpha(\pi - \alpha)}\left(\sin\alpha\sin\theta + \frac{\sin 2\alpha\sin 2\theta}{4} + \frac{\sin 3\alpha\sin 3\theta}{9} + \frac{\sin 4\alpha\sin 4\theta}{16} + \ldots + \frac{\sin n\alpha\sin n\theta}{n^2} + \ldots\right)$$

the summation in this case being made for every integral value of n from one to infinity.

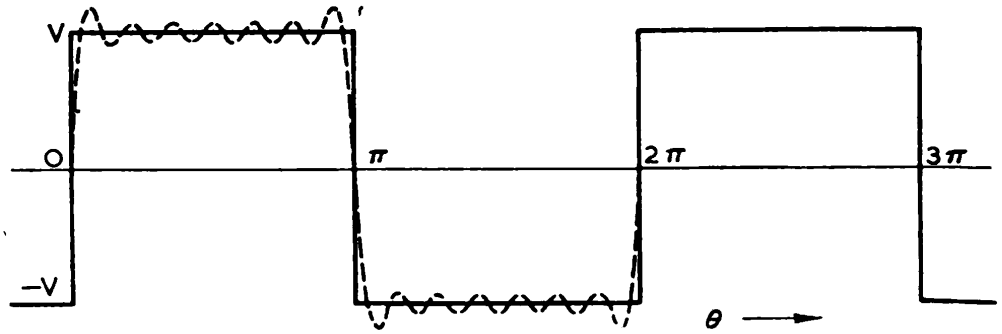

FIG. 4.3.—SQUARE WAVE.

Dotted line shows effect of eliminating all harmonics higher than the 13th.

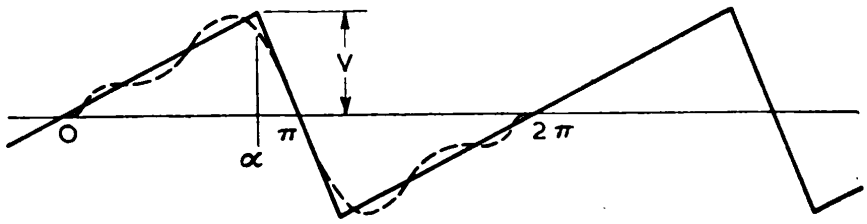

FIG. 4.4.—SAWTOOTH WAVE.

Dotted line shows effect of eliminating all harmonics higher than the 3rd.

Although the above expressions contain an infinite number of terms, the errors introduced in practice are not great if only a reasonable number are taken. The dotted curves in Figs. 4.3

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and 4.4 show the results obtained by neglecting all harmonics after the 13th in the case of the square wave, and the 3rd in the case of the sawtooth. It will be observed that more harmonics are required to give reasonable fidelity in the former case. The practical importance of this analysis is primarily the information it gives regarding the frequency bandwidth with which an amplifier or other circuit must be able to deal if it is to pass waveforms with reasonable fidelity. Truly square waves cannot exist in practice because exact squareness implies an infinite rate of change of current or voltage at points of reversal; nevertheless, a very fair approximation to a square wave can be obtained by taking proper precautions.

Single (Non-recurrent) Pulse

An extension of the foregoing analysis, known as the Fourier integral, shows that a single non-recurrent pulse contains an infinite number of frequencies spaced infinitesimally close together.

A single square pulse of duration T contains all frequencies from zero to infinity, the relative magnitudes of the various frequency components being $\frac{\sin\theta}{\theta}$ where $\theta = \pi fT$, and f is the frequency in question. This is illustrated in Fig. 4.5. If T is measured in seconds or microseconds, f will be in cycles per second or megacycles per second, respectively.

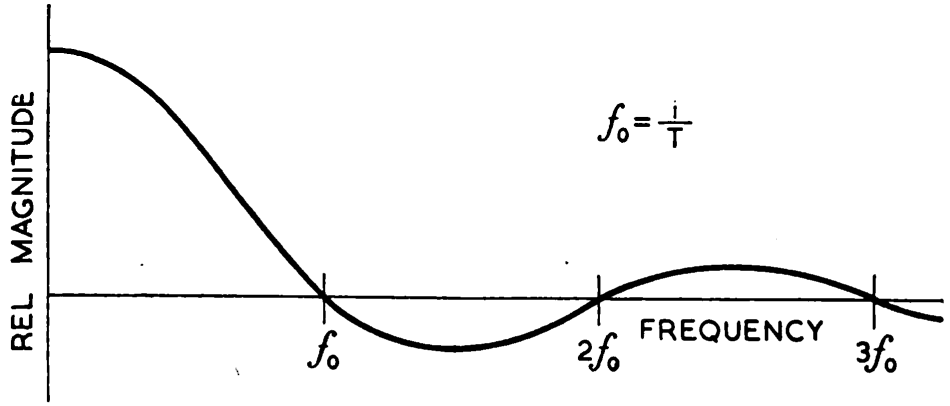

FIG. 4.5.—RELATIVE MAGNITUDES OF THE VARIOUS FREQUENCY COMPONENTS OF A SINGLE PULSE.

It is permissible to use this method for calculating the frequency components of a radar pulse because the pulse duration is so small in comparison with the intervening period between pulses that each may be regarded as an isolated event.

It follows from the above analysis that a radar transmitter consisting of a theoretically perfect oscillator producing "square" pulses will emit sidebands centred on the theoretical frequency of the oscillator and extending, for practical purposes, to frequencies $1/T$ above and below the carrier. This arbitrary limit corresponds to the first zero in Fig. 4.5. Receiver and other circuits must be capable of operating over a bandwidth of this

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amount in order that the pulse shape may be reasonably well preserved. See also Fig. 3.12.

Oscillatory Circuits

The circuit shown in Fig. 4.6 will oscillate, if excited, provided the resistance is not too great.

Suppose the capacitor C is charged to a voltage V immediately before the circuit is completed; the voltage across C at any time t from the moment of closing the circuit will be

$$v = Ve^{-\alpha t} \cos \omega t \text{ where } \omega = 2\pi f \text{ and } f = \frac{1}{2\pi\sqrt{LC}}\sqrt{1 - \frac{R^2C}{4L}}.$$

f is the natural frequency of the circuit.

α is called the damping constant or decay coefficient of the circuit, and its value is $\alpha = \frac{R}{2L}$.

The quantity $\frac{\alpha}{f}$ is called the logarithmic decrement; it is the logarithm (to base e) of the ratio of successive maxima of voltage in the same direction.

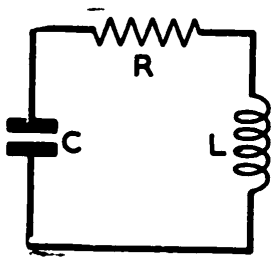

FIG. 4.6.—SIMPLE OSCILLATORY CIRCUIT.

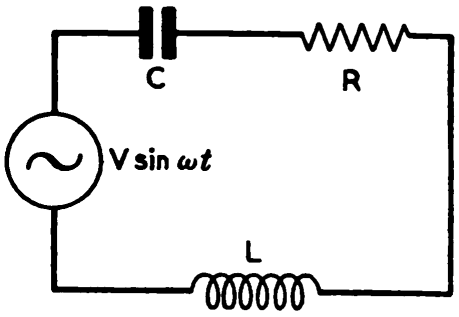

FIG. 4.7.—SERIES RESONANT CIRCUIT.

From the expression for frequency it will be seen that this becomes "imaginary", i.e. the circuit will not oscillate, unless R^2 is less than $4L/C$. If R is equal to this limiting value, the circuit is said to be critically damped.

If the circuit is permanently connected to a source of alternating voltage, as shown in Fig. 4.7, and the supply frequency is adjusted to be equal to $\frac{1}{2\pi\sqrt{LC}}$, the reactance of the inductance will be equal and opposite to that of the capacitance, and the total impedance of the circuit will be simply the resistance R. The current will be $(V/R) \sin \omega t$.

The voltage across the inductance due to this current will be $V(\omega L/R) \cos \omega t$, whence the ratio of the voltage across the inductance to that applied to the whole circuit is $\omega L/R$. In a practical circuit it is impossible to have inductance without some resistance, and the ratio $\omega L/R$ is therefore a measure of the

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"goodness" of the circuit, and is called the "magnification", usually denoted by Q.

The frequency $\frac{1}{2\pi\sqrt{LC}}$ is called the resonant frequency of the circuit. The natural frequency is somewhat lower, but the difference is small with small damping.

Series and Parallel Damping

In addition to the damping produced by the resistance of an inductor (which may be considered as being in series with it), a circuit may in practice be damped additionally by some external load in parallel, as in Fig. 4.8 (a). If the damping is small, this circuit is practically equivalent to that of Fig. 4.8 (b), provided $Rr = \omega^2 L^2$. [R in (b) is simply the equivalent of r in (a); any internal resistance in L must be added to R.]

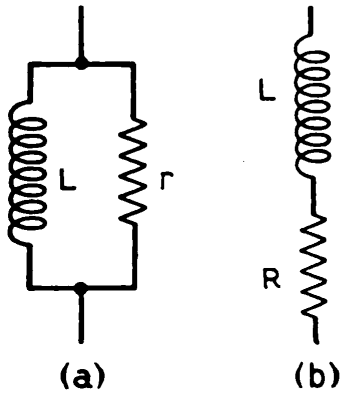

FIG. 4.8.—PARALLEL AND SERIES DAMPING.

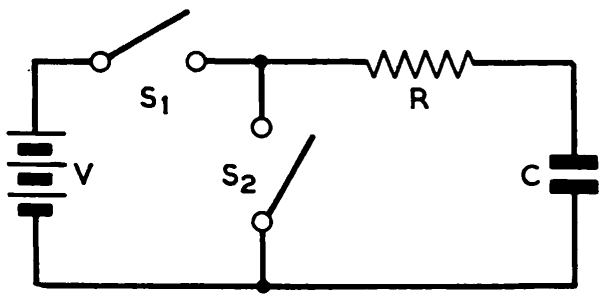

FIG. 4.9.—CHARGE AND DISCHARGE OF CAPACITOR THROUGH RESISTOR.

Charge and Discharge of a Capacitor through a Resistor

In the circuit shown in Fig. 4.9, assume C is initially uncharged, switch S_2 is open, and that S_1 is closed at time $t = 0$. The current flowing through R and C at time t thereafter is $i = \frac{V}{R}e^{-t/RC}$. The voltage across R is therefore $Ve^{-t/RC}$ and across C it is $V(1 - e^{-t/RC})$. These voltages plotted against time are shown in Fig. 4.10, wherein curve A represents the voltage across C and curve B that across R. V is arbitrarily assumed to be 100. The unit of time used in plotting these curves is RC, where the product RC is called the "time constant" of the circuit.

The curves show that at a time equal to the time constant, the current will have fallen to $1/e$ (36·8 per cent) of its initial value, whereas the voltage across C will have risen to 63·2 per cent of the applied voltage.

At a time $t = 5RC$ the current will have fallen to 0·7 per cent and C will be over 99 per cent fully charged; for practical purposes,

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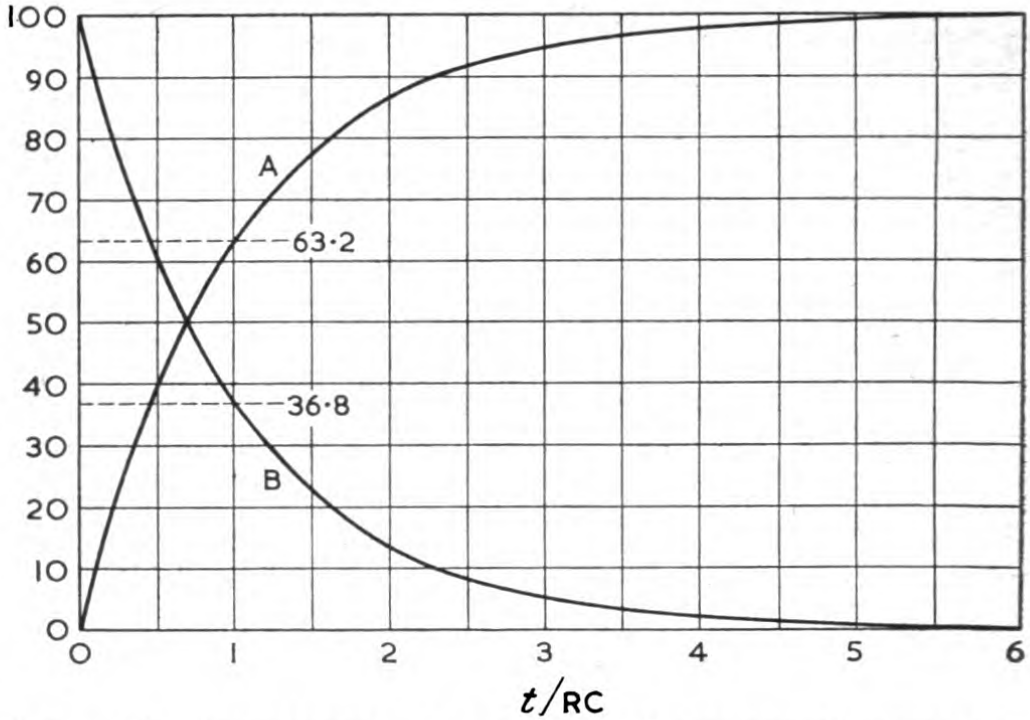

FIG. 4.10.—EXPONENTIAL CHARGE AND DISCHARGE CURVES FOR R–C CIRCUIT.

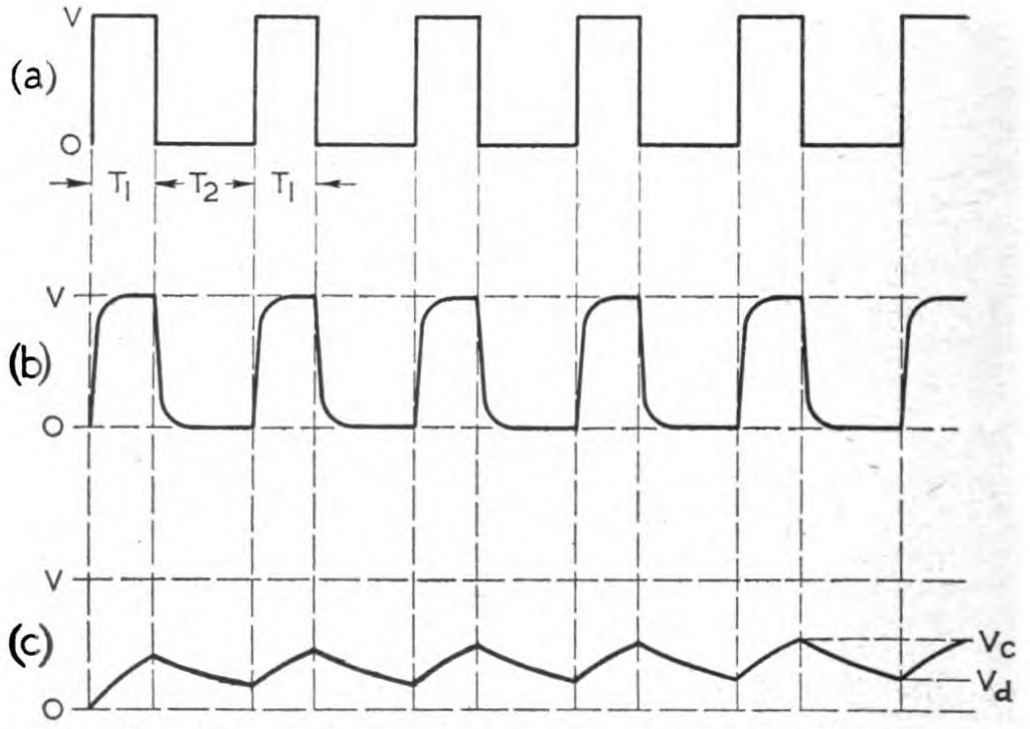

FIG. 4.11.—SQUARE WAVE OF VOLTAGE APPLIED TO R–C CIRCUIT.

(*a*) Applied voltage.
(*b*) Voltage across C for short time-constant circuit.
(*c*) Voltage across C for long time-constant circuit.

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therefore, the process may be considered complete after a lapse of time equal to, say, five times the time constant.

If, after C has become fully charged, switch S_1 is opened and S_2 is closed, the capacitor will discharge through the resistor. During this process, the voltage across C will be represented by curve B, whilst that across R will be equal to it, but with reversed sign. Once again, the process may be considered complete after a lapse of, say, $5RC$.

Current in Circuit containing an Inductor and a Resistor in Series

If C, in Fig. 4.9, is replaced by an inductor L, the initial current, when S_1 is closed, will be zero, but will start to rise according to the law $i = (V/R)(1 - e^{-(R/L)t})$, and will reach 63·2 per cent of its final value after time $t = L/R$, which is the time constant of a circuit consisting of resistance R and inductance L. In this case, the rise of current is represented by curve A in Fig. 4.10, and the voltage across L by curve B.

If, when a steady state has been reached, say in practice after a time $t = 5L/R$, S_1 is opened and S_2 is closed, the well-known inductive effect will cause current to continue to flow in L, but its magnitude will decrease exponentially, i.e. in accordance with curve B.

Circuit Calculations

Many radar circuits have to deal with square waves, and calculations concerning them can frequently be made by judicious application of the facts stated above. The labour involved is in most cases very much less than would be the case if an attempt were made to analyse the waveforms into Fourier series and deal with them as ordinary a.c. problems.

Suppose a square voltage wave, Fig. 4.11 (*a*), is applied to a circuit consisting of a resistance R in series with a capacitance C. If the times for which the voltage is on and off are reasonably large compared with the time constant of the circuit, each part of the cycle can be calculated separately, and the curve of voltage across C, Fig. 4.11 (*b*), obtained. (In the particular case drawn, T_1 and T_2 are 10 and 15 times the time constant, respectively.)

If the time constant is too long for the above method to be valid, the capacitor will neither charge nor discharge fully during each cycle, but a steady state will soon be reached, Fig. 4.11 (*c*), wherein the limits of voltage across C are, say, V and V_c. Once the values of these voltages are known, the calculations can be made by starting at the appropriate points on the curves of Fig. 4.10.

It can be shown mathematically that $V_c = V\dfrac{(1-p)}{(1-pq)}$ and $V_d = V\dfrac{q(1-p)}{(1-pq)}$, where $p = e^{-T_1/RC}$ and $q = e^{-T_2/RC}$.

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Alternatively, the values of V_c and V_d can be found by trial and error, using the curves of Fig. 4.10, the conditions being that the voltage must rise from V_d to V_c in the charging time T_1 and fall from V_c to V_d in the discharging time T_2.

A circuit consisting of inductance and resistance can be dealt with in a generally similar manner.

Differentiation

The current (I amperes) flowing into a capacitor (C farads) is equal to the product of the capacitance C and the rate of change of voltage across it. Expressed mathematically, $I = C\frac{dV}{dt}$.

If, in the circuit of Fig. 4.12, the applied voltage V_1 is constant, no current will flow, after the initial transient, and $V_3 = V_1$, $V_2 = 0$. If, however, V_1 changes, and the time constant RC is short,* V_3 will follow the change in V_1 rapidly, the voltage drop (V_2) across R being small. V_3 is therefore very nearly equal to V_1 and the current in C is therefore very nearly pro-

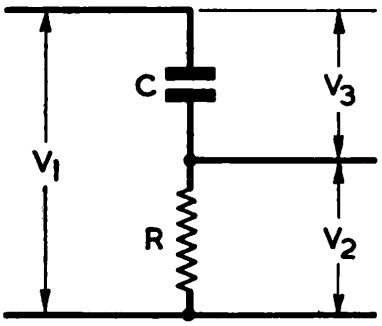

FIG. 4.12.—R–C DIFFERENTIATING CIRCUIT.

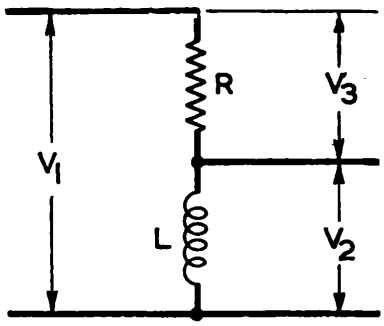

FIG. 4.13.—L–C DIFFERENTIATING CIRCUIT.

portional to the rate of change of V_1. Since the drop across R is proportional to the current, it follows that V_2 is proportional to the rate of change of V_1. Expressed mathematically, $V_2 = RC\frac{dV_1}{dt}$.

The mathematical operation of finding a quantity proportional to the rate of change of a given quantity is called "differentiation", whence this type of circuit is called a differentiating circuit.

The circuit shown in Fig. 4.13 is also a differentiating circuit provided the time constant L/R is short. In this case we have $V_2 = (L/R)\frac{dV_1}{dt}$.

Uses of Differentiating Circuits

These are used frequently for producing sharp peaky waveforms from square waves. Fig. 4.14 shows the voltages across

* "Short" in this connection means short in comparison with the time occupied by the voltage change under consideration.

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C and R when a square voltage wave is applied to the circuit of Fig. 4.12, the time constant RC being small compared with the period of the applied voltage.

The peaky output wave is used for many purposes in radar, e.g. triggering, producing calibration pips, etc. Frequently only the positive going pips are required, in which case the others can be eliminated by passing through a valve biased beyond cut-off in one direction. If the dotted line in the V_2 curve of Fig. 4.14 represents cut-off, only those parts of the wave above this line are retained. This expedient narrows the base of the pulse, thus sharpening it.

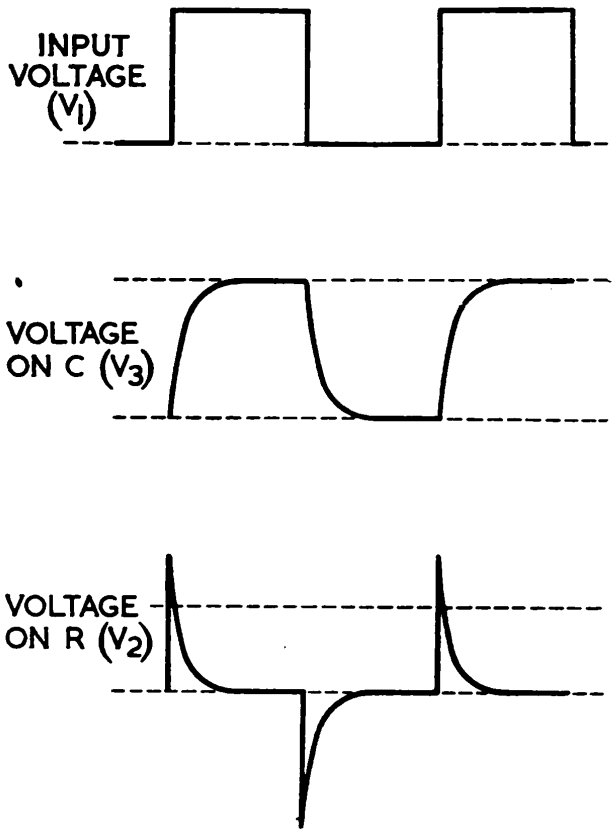

FIG. 4.14.—PRODUCTION OF PEAKY WAVEFORM BY DIFFERENTIATING SQUARE WAVE.

Integrating Circuits

Integration is the reverse of differentiation. If, in the circuits shown in Figs. 4.12 and 4.13, the values of R, C and L are chosen to make the time constants long, and if V_3 (instead of V_2) is taken as the output, the rate of change of the output will be proportional to the input. Expressed mathematically,

$$V_3 = 1/RC \int V_1 dt \text{ for Fig. 4.12 and } V_3 = R/L \int V_1 dt \text{ for Fig. 4.13.}$$

(These relations are very nearly correct, provided the time constant is long.)

Thévenin's Theorem

This is frequently useful in calculating currents and voltages in networks containing only linear elements. A linear element is one whose impedance does not alter when the current through it changes. It may be a resistance, inductance, capacitance, or a combination of these.

The theorem may be stated as follows. Suppose a linear network has four terminals, and that a generator connected to terminals 1 and 2 produces a voltage V between terminals 3

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and 4. Now suppose that a load of impedance Z_1 is connected between terminals 3 and 4, the current which will flow in this load is $\frac{V}{Z_1 + Z_2}$, where Z_2 is the impedance of the network measured between terminals 3 and 4 with the generator across terminals 1 and 2 replaced by a passive impedance equal to the internal impedance of the generator.

Reciprocity Theorem

Suppose a generator is connected between terminals 1 and 2 of the above network and an ammeter across 3 and 4. If the impedances of generator and ammeter are equal, the current shown will be the same if the positions of generator and ammeter are interchanged.

If generator and ammeter each have zero impedance, the ratio $\frac{\text{input voltage}}{\text{output current}}$ is called the transfer impedance of the network.

Resistance of Conductors Carrying A.C. (Skin Effect)

When a conductor is carrying A.C., eddy currents are induced which affect the distribution of the current over the cross-section of the conductor. The current density is greatest at the outside of the conductor, and decreases towards its interior ; causing the effective a.c. resistance to be greater than the d.c. value.

For the very high frequencies used in radar, the effective resistance can be calculated to a fair degree of accuracy by assuming that the whole of the current is confined to a thin layer at the surface of the conductor. The thickness of this layer (skin depth) is, in M.K.S. units, $\sqrt{\frac{\rho}{\pi\mu f}}$, where ρ is the specific resistance of the conductor material, μ its absolute permeability, and f the frequency in cycles per second.

For copper at 20° C. this becomes $\frac{0{\cdot}066}{\sqrt{F}}$ millimetre, where F is the frequency in megacycles per second.

Heating Effect of Pulse Currents

When designing those parts of a radar circuit which carry pulse currents of appreciable magnitude, regard must be had to the fact that the heating effect in a conductor is proportional to the product of the m.s. value (Mean square, equal to r.m.s. value squared) of the current and the resistance.

For a sinusoidal waveform, the r.m.s. value is 1·11 times the mean (this is the form factor) but for a square wave which is

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" on " for only $1/n$ of the total time, the figure is $\sqrt{n}$. Since n is often of the order of 1,000, the heating effect, if calculated from the mean value of the current, as shown by a moving-coil meter, would be underestimated a thousand times.

This also shows why moving-coil meters in pulsed circuits must be protected by suitable capacitors across them if heavy currents are involved.

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CHAPTER V

SQUARE WAVEFORMS AND PULSES FOR CONTROL PURPOSES

THE frequency with which the complete cycle of events is repeated, i.e. the pulse repetition frequency, may be controlled in any of the following ways :

(*a*) By the frequency of the a.c. power supply, if suitable.
(*b*) Master oscillator.
(*c*) Free-running multivibrator.
(*d*) Crystal-controlled oscillator, in special cases.

Synchronizing

The circuit or unit controlling the p.r.f. may also directly control other circuits, e.g. timebase, etc., or these may be triggered by a pulse from the transmitter circuit.

It is essential that all the various units in an installation be properly synchronized ; this applies especially to the timebase and calibrator circuits. Other circuits, e.g. the c.r.t. brightening pulse, swept gain, etc., must also be synchronized, but exact synchronism is not so important in these cases.

Whatever the prime controlling source, it is usual to develop square switching waves by means of controlled multivibrators, usually having one stable and one quasi-stable state. These switching waves are then used for operating the various circuits. Reversion from the quasi-stable to the stable state usually occurs a short time after the completion of the range timebase sweep, in sufficient time for everything to be in readiness for the next cycle. Subject to this, the exact time of reversion is not usually important.

Other methods may be used in special cases, as for example the coherent pulse technique for moving target indication described in Chapter XIII.

Multivibrator

This device, which may be either free-running or triggered, is extensively used for the generation of square waveforms.

Fig. 5.1 shows a typical circuit. This is virtually a two-stage amplifier with the output coupled back to the input. In operation, as a result of the heavy feedback, the two valves are alternately driven beyond cut-off and then into full conduction.

Suppose that V_1 has just been driven to cut-off. V_2 will be fully conducting, and its anode voltage will be below that of the H.T. supply owing to drop in R_2. It remains at this point until, due to discharge of C_2 through R_3, the grid potential of

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V_1 rises sufficiently for this valve to start to conduct. The anode voltage of V_1 then starts to fall, and with it the grid of V_2. The process is cumulative, and V_2 is rapidly driven to

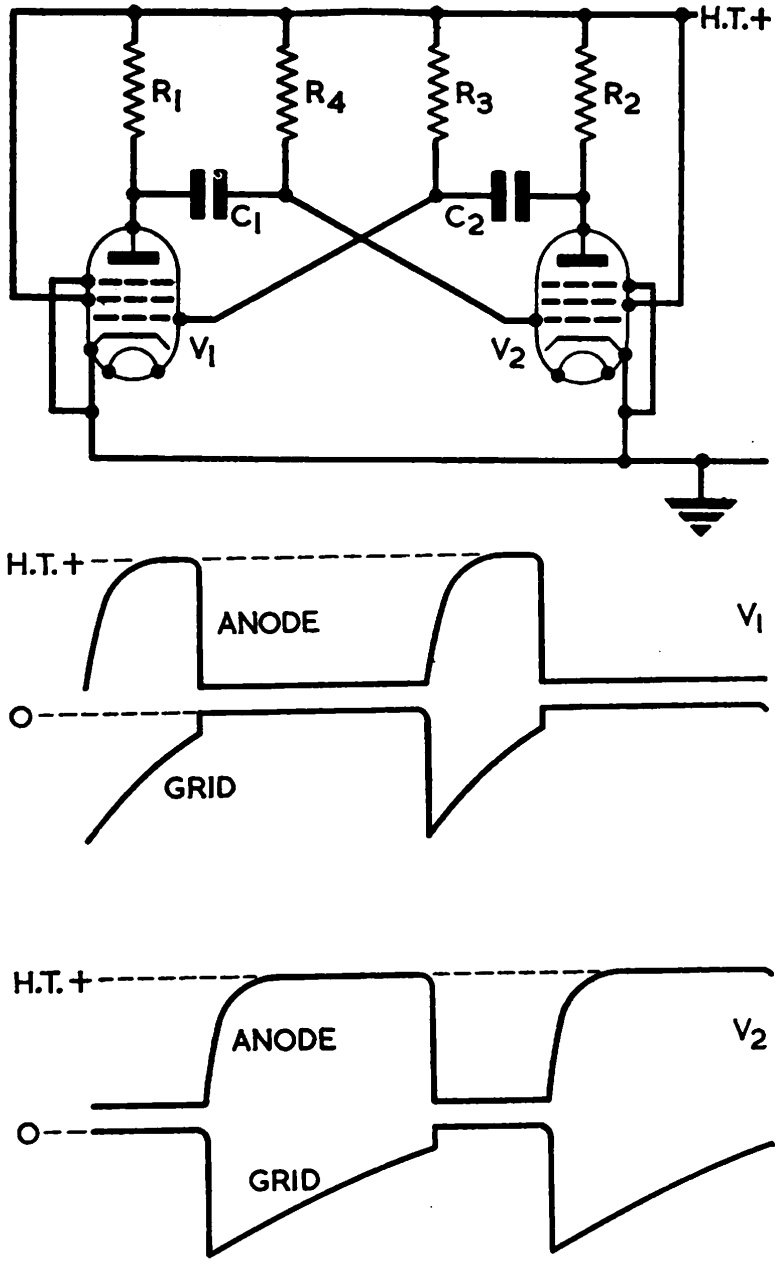

FIG. 5.1.—FREE-RUNNING MULTIVIBRATOR.

cut-off. This change-over takes place practically instantaneously provided stray capacitances are not too great. Conditions are now as at the start, except that the roles of V_1 and V_2 are interchanged.

The same process is repeated until V_2 again conducts and V_1 is cut-off, and so on, indefinitely. The result is the production of square (nearly) waveforms at the two anodes, in

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antiphase to one another, and of durations primarily depending on the time constants of R_4C_1 and R_3C_2. These are shown in Fig. 5.1.

The grid leaks R_3 and R_4 may be connected to earth or to H.T.+ as shown. The latter is often preferred as it makes the change-over more rapid.

The repetition frequency of the free-running multivibrator is not of a high order of stability, but this is unimportant in many cases. If desired, the frequency can be stabilized by injecting into one of the grid circuits a sinusoidal voltage from a stable oscillator. The natural frequency of the multivibrator must be sufficiently near to that of the oscillator, or a multiple or sub-multiple of it, to enable it to lock on to the injected frequency. This technique is of considerable importance in the accurate measurement of frequency, where the control oscillator may be a crystal, but has no other particular application to practical radar.

An alternative method of stabilizing is to use a triggered multivibrator controlled by accurately spaced trigger pulses.

Triggered Multivibrator or Flip-Flop

Fig. 5.2 shows a cathode coupled multivibrator in which V_2 is normally conducting and V_1 is cut off by bias due to the anode

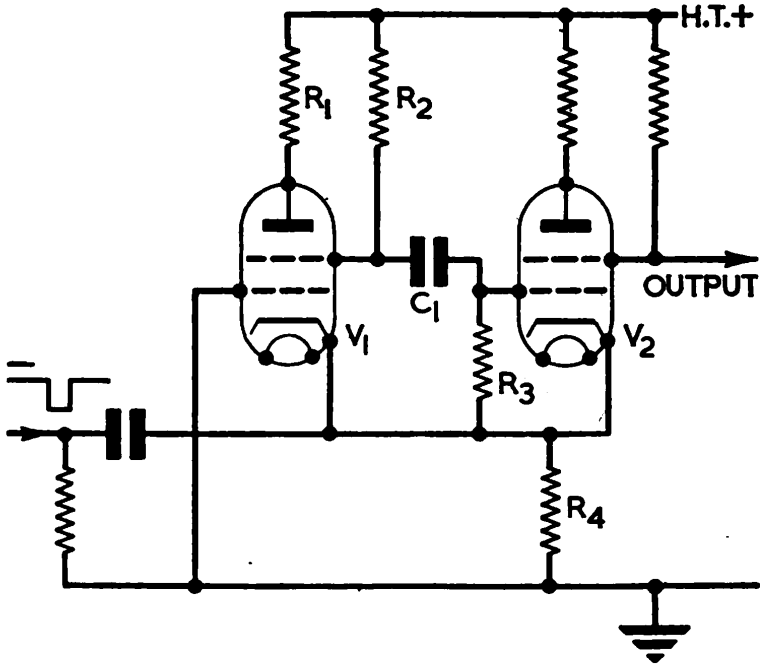

FIG. 5.2.—TRIGGERED MULTIVIBRATOR OR FLIP-FLOP.

current of V_2 flowing in R_4. Unless disturbed, the circuit will remain in this condition. If a negative trigger pulse is applied to the common cathode circuit, the bias across R_4 will be neutralized, V_1 will conduct and drive V_2 to cut off. After a time

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determined by the time constant of R_3C_1 the circuit will revert to its original state, and remain so until another trigger pulse arrives. This multivibrator produces an approximately square

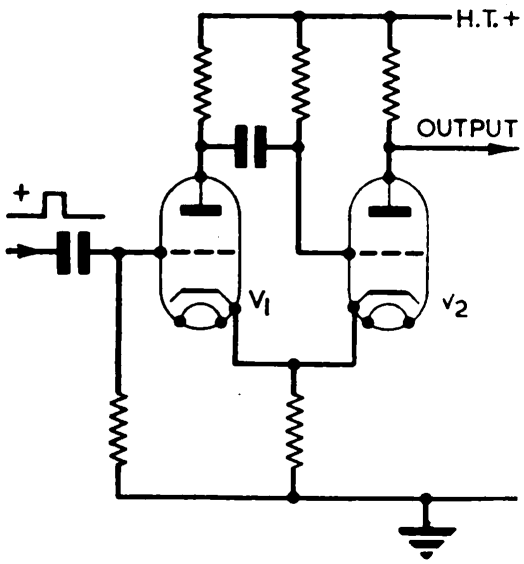

FIG. 5.3.—CATHODE-COUPLED MULTIVIBRATOR.

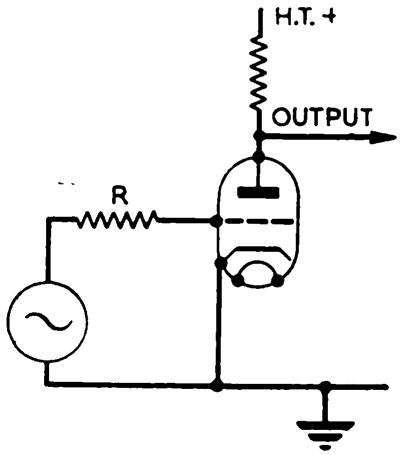

FIG. 5.4.—GENERATION OF SQUARE WAVES FROM SINUSOIDAL A.C.

waveform, as in the case of the free-running type, but each cycle has to be triggered, the device not being self-repetitive.

Another example is shown in Fig. 5.3. In this circuit V_1 is cut off until the application of a positive trigger pulse drives it to conduction and V_2 to cut off.

Delay Circuits

When it is required to introduce a delay into a circuit, e.g. to produce a trigger pulse at some specified interval after the application of an initiating trigger pulse, the triggered multivibrator may be used, its recovery time being adjusted to the required time interval, and the swing of anode voltage on reversion used to produce the required second trigger.

More accurate methods employ artificial delay lines, the initiating trigger

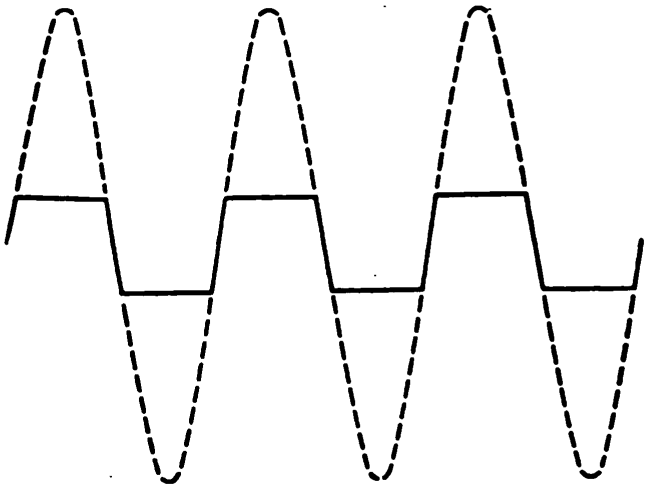

FIG. 5.5.—WAVEFORM OBTAINED FROM CIRCUIT OF FIG. 5.4.

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pulse being fed into one end of the line and received, after the appropriate lapse of time, at the other.

Another method is to trigger an accurate timebase circuit and cause another trigger to be produced when the timebase voltage reaches some particular value. Yet another method is to con-

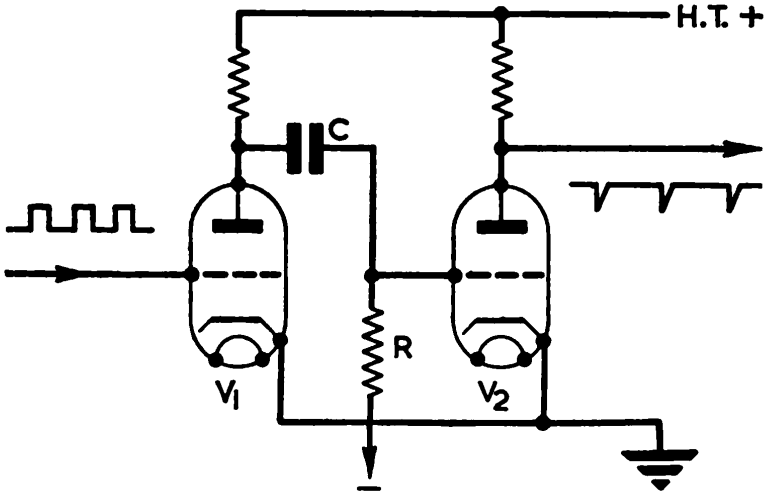

FIG. 5.6.—PRODUCTION OF SHORT PULSES.

vert the electrical trigger pulse into a supersonic sound wave by means of a piezo crystal, pass the sound wave through a tube filled with mercury, along which it travels with the speed of sound, and convert it back to an electrical impulse with another crystal. This method is used for rather special applications, and is not usually found in a normal radar installation. In some cases, the sound wave may be reflected from the far end of the tube, and the same crystal used for both purposes.

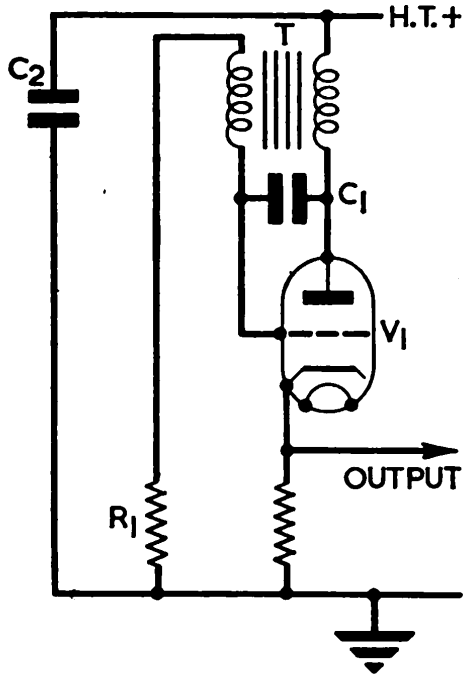

FIG. 5.7.—BLOCKING OSCILLATOR.

Generation of Square Waves from Sinusoidal A.C.

The circuit shown in Fig. 5.4, if overdriven by the application of sufficient a.c. voltage to the grid of the valve, will produce at the anode a voltage waveform which is limited on its positive-going excursion by the valve cutting off, and in the other direction by grid current in *R* preventing appreciable rise of grid voltage above that of the cathode. The output waveform is shown by the full line in Fig. 5.5. The " vertical " sides can be made as steep as desired,

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within reason, by making the input voltage sufficiently large. The frequency is of course that of the a.c. supply.

Trigger Pulses

Short pulses for triggering multivibrators and other circuits can be derived from square waves by various means. One example is shown in Fig. 5.6. A square wave is fed to the grid of V_1 and the output is differentiated by the R–C circuit, which has a small time constant. V_2 is biased beyond cut-off so that only the positive tips of the waveform at its grid produce anode current—see Fig. 4.14. The output of V_2 therefore consists of a series of very short pulses which can be used to trigger other circuits.

The Blocking Oscillator

This circuit, shown in Fig. 5.7, can be used for producing short sharp pulses.

The transformer T gives positive feedback of much greater

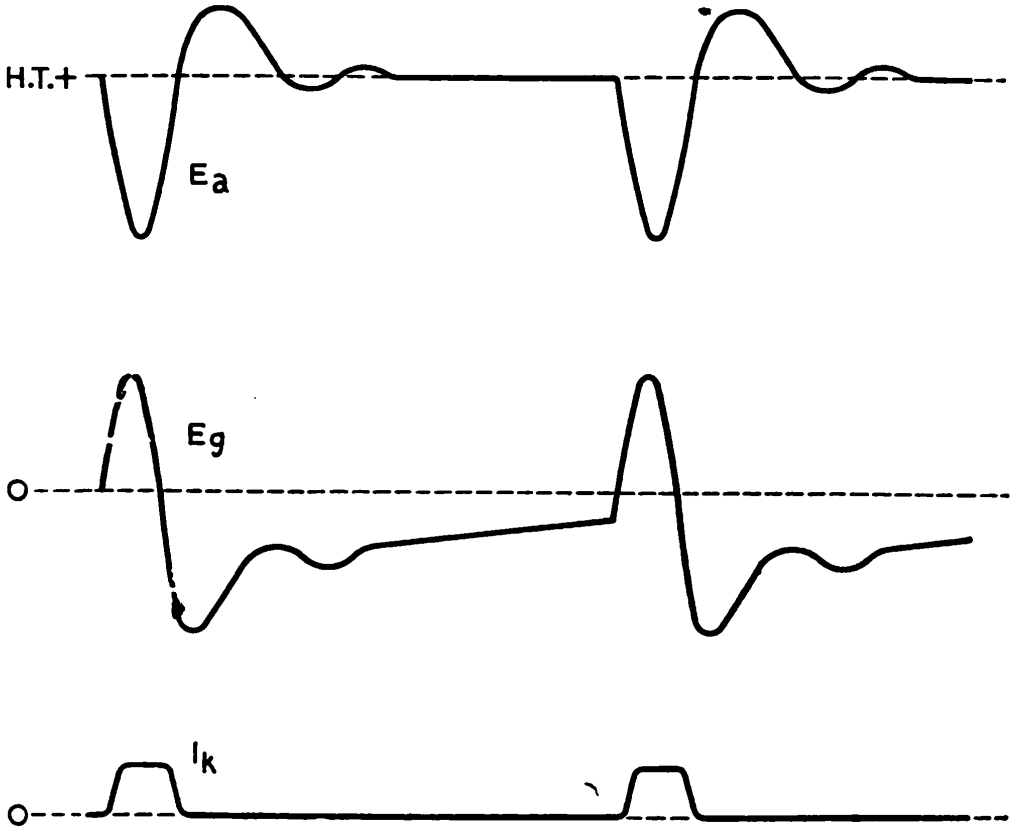

FIG. 5.8.—BLOCKING OSCILLATOR WAVEFORMS.

magnitude than in a normal oscillator. C_1 is small, consisting mainly of the interwinding capacitance of T plus strays; C_2 is large. The windings of T, together with C_1, form an oscillatory circuit of high natural frequency.

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When the circuit is switched on, anode current flows, the grid is driven strongly positive, and grid current charges C_1 to a sufficient voltage for V_1 to be cut off on the negative half-cycle of oscillation. The valve being cut off, there is no further feedback, and subsequent cycles of oscillation are rapidly damped out, largely by iron losses in the transformer. Typical waveforms of anode and grid voltage and cathode current are shown in Fig. 5.8.

When the charge on C_1 has leaked away through R sufficiently for V_1 to conduct again, the cycle is repeated.

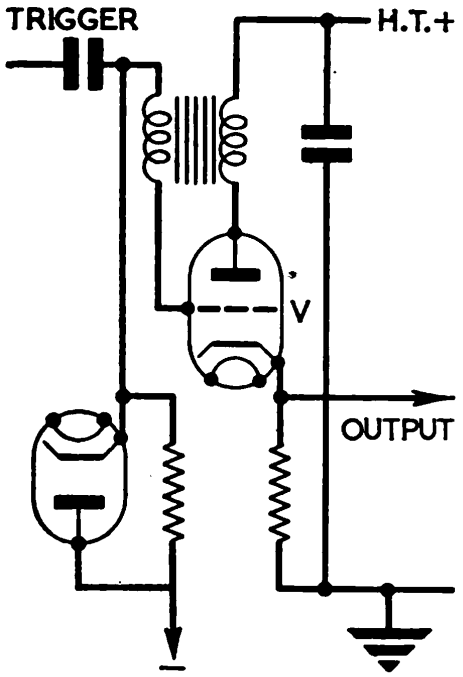

FIG. 5.9.—TRIGGERED BLOCKING OSCILLATOR.

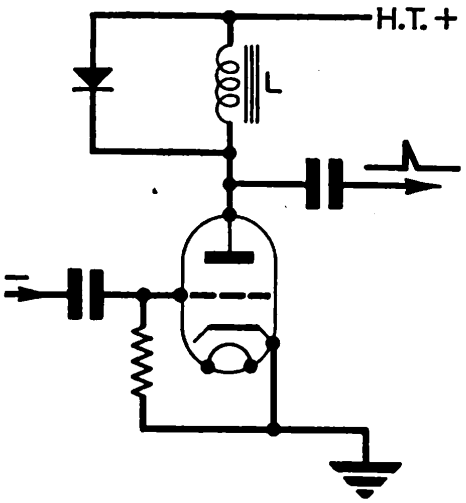

FIG. 5.10.—DERIVATION OF PULSE FROM INDUCTIVE KICK.

The blocking oscillator is frequently used for controlling the whole cycle of operations in a radar installation, the pulses it produces being used for triggering other circuits.

Triggered Blocking Oscillator

The output pulse from a blocking oscillator can be made very short, and the circuit is often used for the production of calibrator pips, etc., where a very short pulse is required. In such cases the circuit must be triggered rather than free-running, the necessary modifications being shown in Fig. 5.9.

External bias sufficient to cut off V without the aid of extra bias generated by grid rectification is required. The diode prevents the grid going too negative.

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Pulse Produced by Opening Inductive Circuit

The circuit shown in Fig. 5.10 will produce a positive-going pulse if the current in the valve is suddenly cut off by the application of a relatively large negative voltage to its grid. A germanium crystal connected across the inductance suppresses any tendency to further oscillation. The pulses produced by this circuit can be made sharp enough for use as calibration pips.

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CHAPTER VI

RADAR TRANSMITTERS

THE main parts of a radar transmitter are, (1) the oscillation generator, (2) the modulator for controlling it, (3) power amplifier for amplifying the oscillator output (only applicable to older transmitters working at metre wavelengths), (4) feeder and aerial system, (5) aerial turning mechanism, (6) power supply. Only the first three of these are dealt with in this chapter, the others being considered later.

In addition, there are various ancillary items such as test gear, monitoring devices, cooling fans, etc.

Pulse Length, Repetition Frequency, Power

Pulse lengths are generally of the order of microseconds, short pulses of from 0·1 to 1 microsecond being usual for accurate work; longer pulses up to about 10 microseconds are used for less accurate work at greater ranges. Pulse repetition frequencies vary from about 200 to 2,000 per second. The ratio

$$\frac{\text{pulse duration}}{\text{total period from start of one pulse to the next}}$$

is called the " duty cycle " and in most installations lies between approximately 0·0005 and 0·002.

The *mean* power emitted = peak power × duty cycle. Peak powers in radar are very high compared with those used in conventional communication systems, but mean powers are quite moderate.

Metre Wave Transmitters

These are becoming less common nowadays. They employ valve oscillators, generally operated in push-pull circuits, usually followed by a push-pull amplifying stage feeding the aerial. The tuned circuits are short lengths of transmission line (lecher wires) because, even at these comparatively long wavelengths, circuits consisting of lumped inductance and capacitance are not very practicable.

Centimetre Wave Transmitters

Most modern radars work in the centimetre band, particularly between 10 and 3 centimetres. Except in a few special installations of low power, the magnetron is universally used as the source of oscillation. Some work has been done with high-power klystrons, but so far their application to radar may be regarded as experimental.

The magnetron requires a high voltage d.c. pulse applied to

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it for a period equal to the pulse length required. Basically all that is required is a high-voltage d.c. supply and a modulator to switch it on and off.

The use of a pulse transformer enables the modulator to operate at a somewhat lower voltage than that required by the magnetron; it has also certain advantages from the circuit designer's point of view.

Peak powers up to about 1,000 and 200 kilowatts at 10 and 3 centimetres, respectively, can be obtained.

Generation of Short Square Pulses for Transmitters—Artificial Lines

The pulse from a blocking oscillator is sometimes used to control the modulator of a radar transmitter, but methods using artificial lines are generally preferred. The artificial line can be used to form a pulse of the required duration at high power, subsequent amplification being unnecessary.

Consider a length of transmission line, of characteristic impedance Z_0 (see Chapter VIII), connected as in Fig. 6.1. The line is charged from a d.c. supply (not shown) to a voltage V, and the load, of resistance $R = Z_0$, is then connected by closing the switch S. A current $V(Z_0 + R) = V/2R$ flows in the load, the voltage at the load end of the line immediately falling to $V/2$. So far as the line is concerned, this fall of voltage is equivalent to the application of a voltage $-V/2$ to the load end. This produces a voltage wave $-V/2$ which travels along the line, reducing the voltage at every point along the line, progressively, from V to $V/2$. At the far end the wave is reflected without change of phase, and the returning wave $(-V/2)$ cancels out the remaining voltage on the line. By the time the wave arrives back at the starting-point, the line is completely discharged and the current in R ceases. This circuit therefore produces in R a square pulse of current which lasts for time $T = 2l/v$,

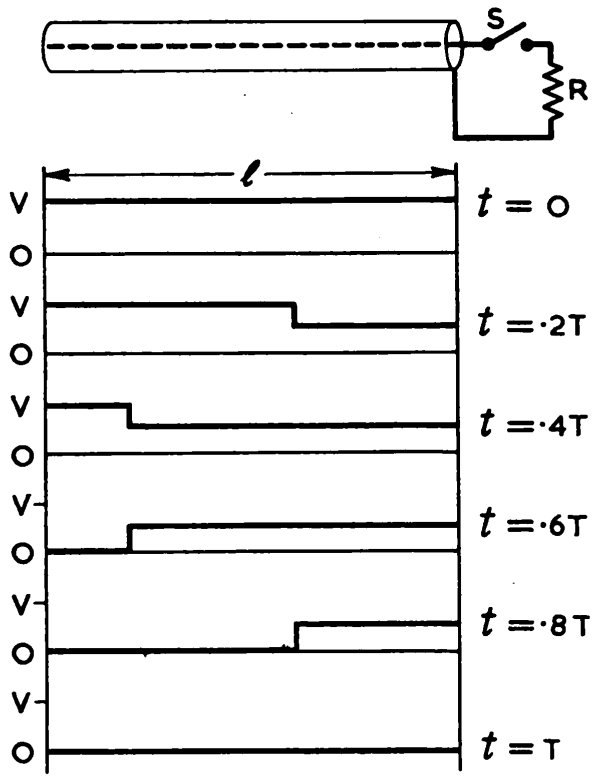

FIG. 6.1.—PRODUCTION OF PULSE BY DISCHARGING TRANSMISSION LINE.

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where l is the length of the line and v the velocity of propagation of the wave along it. The lower part of Fig. 6.1 shows the voltages on the line at various instants.

As an actual line would be inconveniently long (of the order of 100 yards per microsecond), an artificial line, consisting of lumped inductances and capacitances (Fig. 6.2), is used in practice. Artificial lines are not exact equivalents of actual

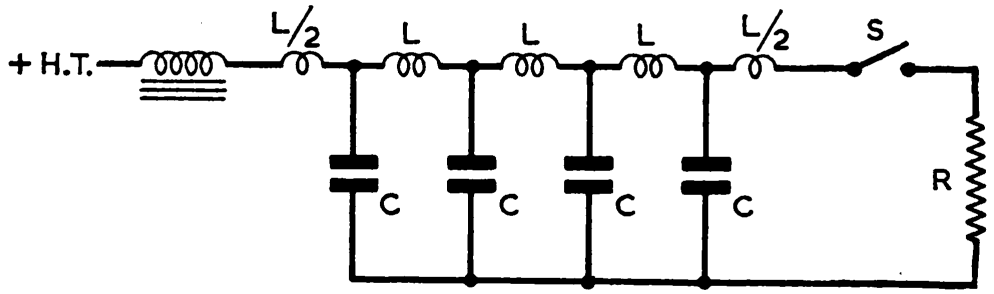

FIG. 6.2.—ARTIFICIAL LINE.

uniform lines, and some distortion of pulse shape results from their use, but this can be reduced to unobjectionable proportions by proper design. For transmitters, an artificial line consisting of three or four L–C sections is usually sufficient.

The switch S may be a thyratron, a triggered spark gap, or a suitable hard valve.

The line can be charged from the H.T. supply through a choke, which will have substantially infinite impedance at frequencies corresponding to the pulse length.

Modulators

An example of a modulator circuit using an artificial line is shown in Fig. 6.3. The " switch " is a thyratron, V_1, fired by the application of a positive trigger pulse to its grid. The magnetron is fed through a pulse transformer which carries a tertiary winding to provide an output trigger pulse, coincident in time with the firing of the magnetron, which can be used for triggering the timebase circuit, or other purposes if required.

As a magnetron is not a load of constant resistance, correct matching to the feeder is theoretically impossible, and a compromise has to be made. Any resulting tendency to overswing and produce a reverse pulse is suppressed by the diode V_2.

Another circuit differing in detail, but not in principle, is shown in Fig. 6.4. In this case, a triggered spark gap, called a " trigatron ", is used as the switch.

In both cases the anode of the magnetron is at earth potential, an important practical point because the output circuit, whether coaxial line or waveguide, is connected to the anode. The high voltage pulse, which is of course negative, is applied to the cathode.

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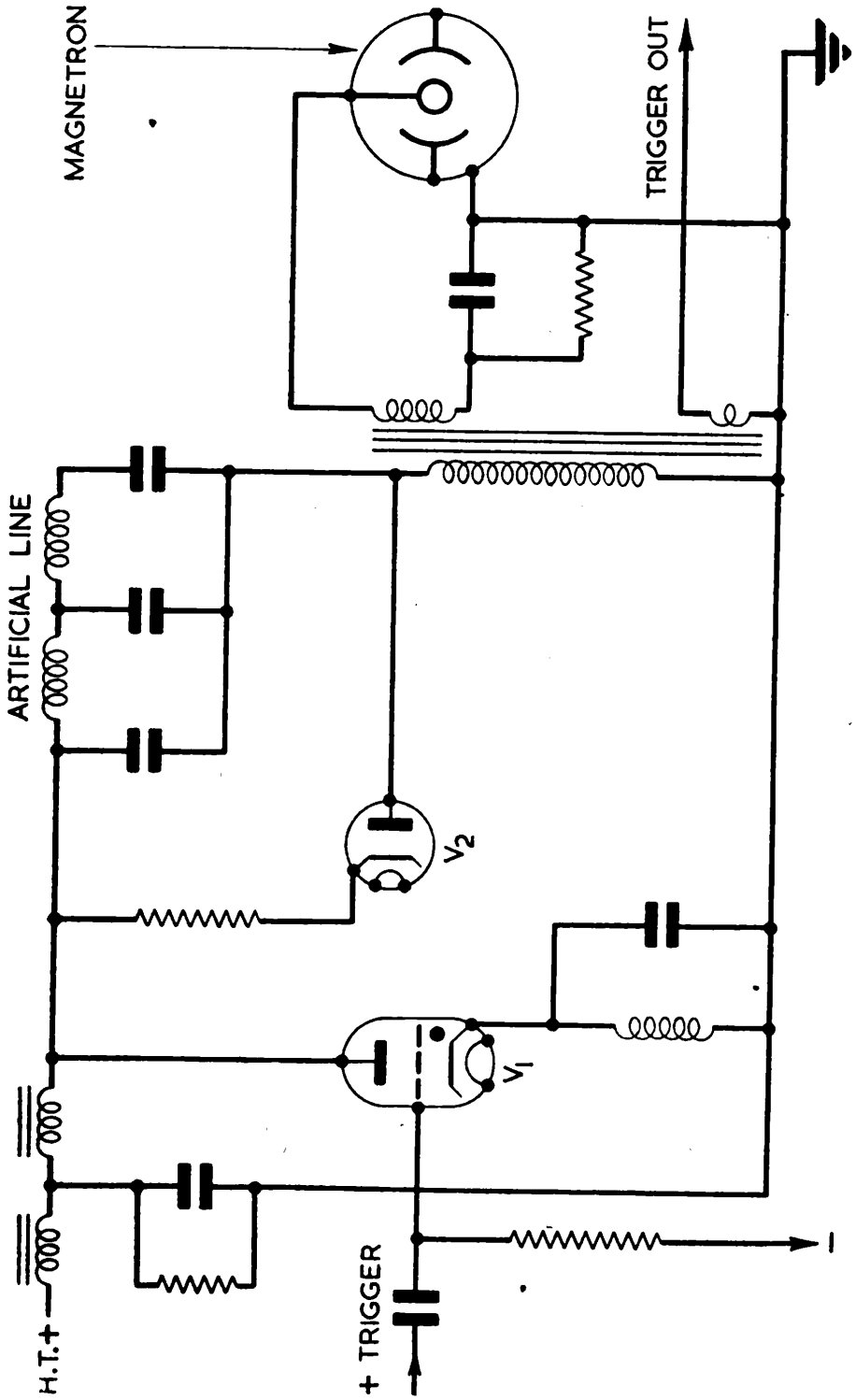

FIG. 6.3.—MODULATOR USING ARTIFICIAL LINE AND THYRATRON.

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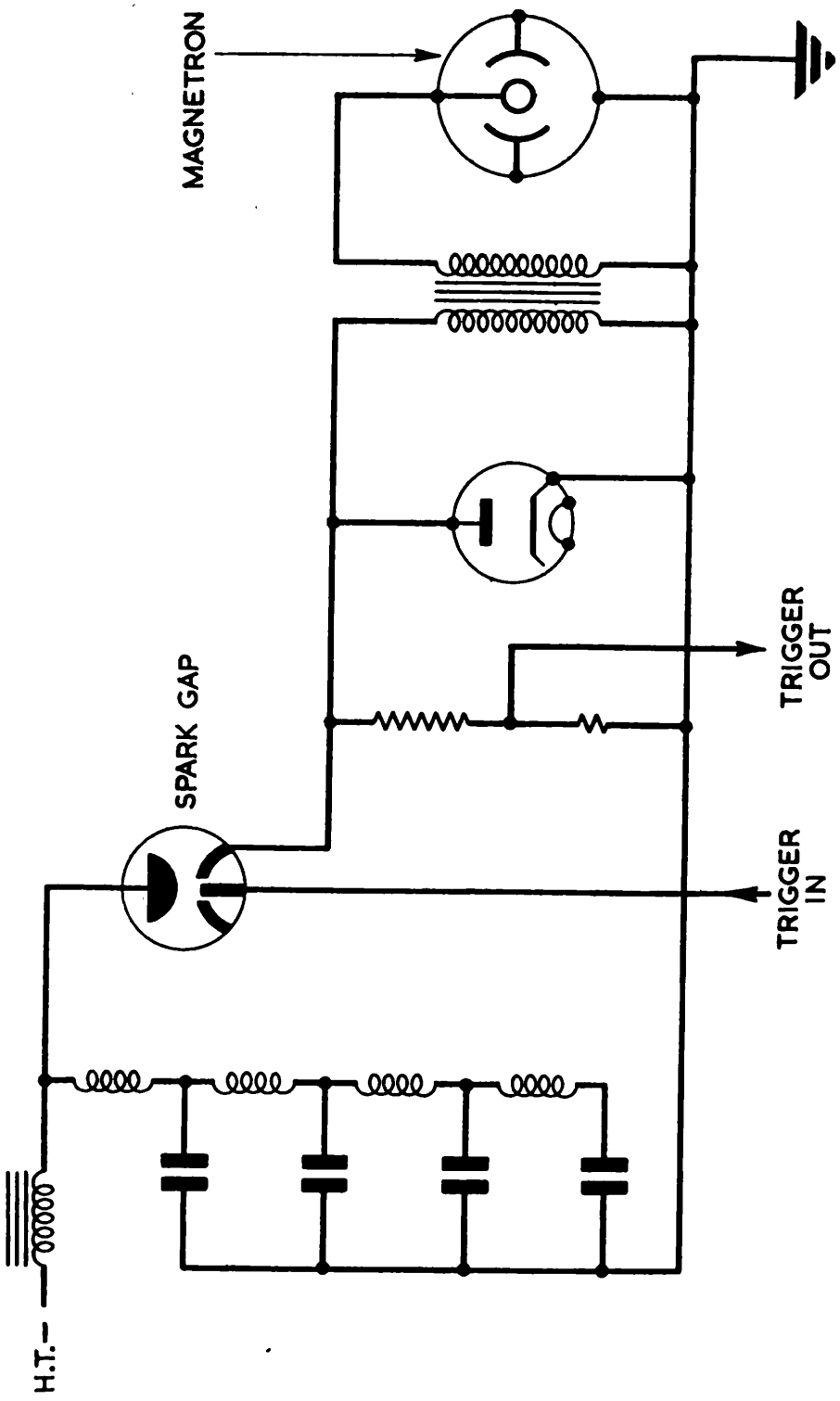

FIG. 6.4.—MODULATOR USING ARTIFICIAL LINE AND SPARK GAP.

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Triggered Spark Gap, or Trigatron

This consists of two hemispherical electrodes, generally of molybdenum (Fig. 6.5). The gap is not broken down by the normal working voltage applied to the main electrodes, but when a trigger pulse is applied to the trigger electrode, the main gap is caused to break down.

The trigatron may be open or enclosed; in the latter case a filling of argon with about 5 per cent oxygen has been found satisfactory. Due to the action of the spark the electrodes require adjustment about every 100 hours, and renewal from time to time.

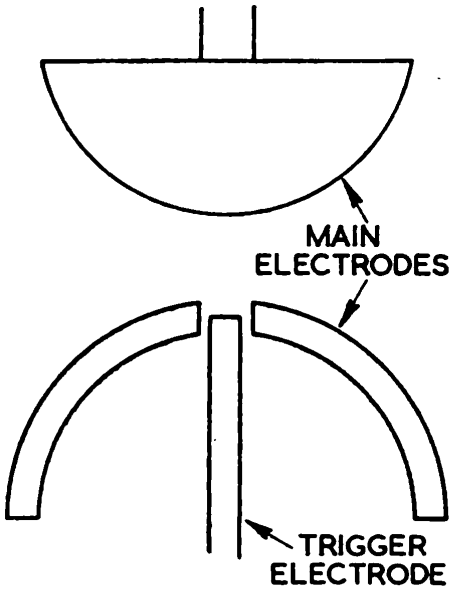

FIG. 6.5.—SPARK GAP.

Pulse Transformer

To reproduce the pulse shape accurately, it is necessary to reduce as far as possible the effects of leakage inductance and stray and interwinding capacitances. Iron losses must be minimized by the use of a suitable core material in very thin laminations.

The secondary winding is frequently wound with two wires in parallel (bifilar winding). These two wires convey the heater current; this obviates the necessity of designing a heater transformer to stand up to the pulse voltage, and of sufficiently low capacitance to avoid loading it.

Resonant Charging

An artificial line can be charged from an a.c. supply without a rectifier by making a choke in the charging circuit resonate,

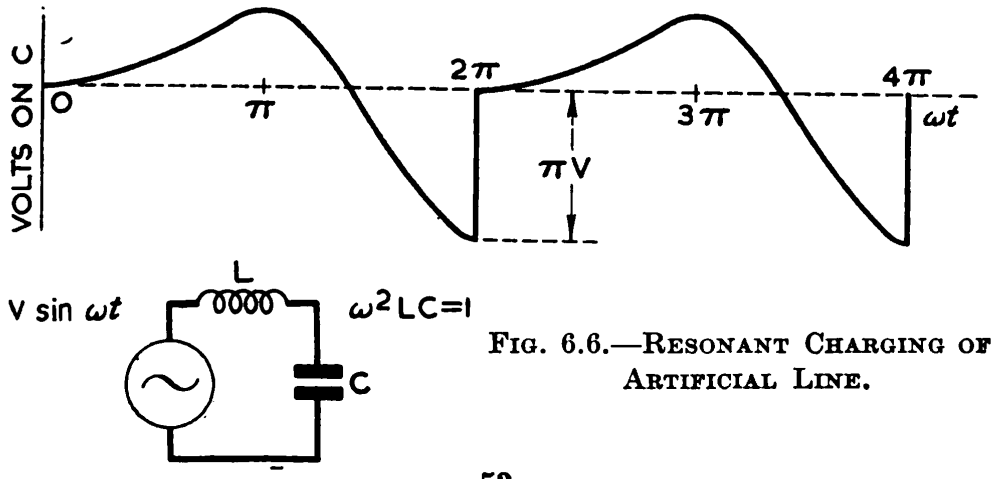

FIG. 6.6.—RESONANT CHARGING OF ARTIFICIAL LINE.

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at the supply frequency, with the capacitance of the line. (See Fig. 6.6.) If C is uncharged at time $t = 0$ when the circuit is switched on, the voltage across C at any subsequent time (neglecting losses) is $\frac{1}{2}V(\sin \omega t - \omega t \cos \omega t)$. If the line is discharged at the end of every complete cycle of the supply frequency, the initial condition (zero voltage across C) is restored, and the process can be repeated indefinitely. It is assumed that the pulse length is so small compared with a cycle of the supply frequency, that the discharge which forms the pulse may be considered instantaneous.

Hard Valve Modulator

A pulse of suitable length from a blocking oscillator or other pulse generating circuit may be used to control the modulator shown in Fig. 6.7. R is the load (magnetron or other oscillator), and the modulator valve V is normally biased beyond cut-off so that it only conducts when a positive pulse of sufficient magnitude is applied to its grid. The capacitor C must be large enough to supply the anode current without appreciable voltage drop during the pulse.

This circuit suffers from the fact that, in general, a high value of H.T. is required, and this necessitates a large grid swing to take the valve from cut-off to a condition of conduction without undue drop in the valve itself. The voltage of the control pulse is, in fact, comparable with that of the final power pulse.

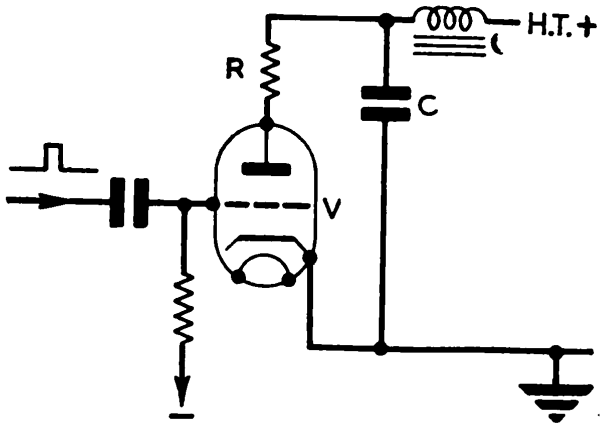

FIG. 6.7.—HARD VALVE MODULATOR.

The difficulty can be overcome by using a modification of the cathode follower circuit, called the " bootstrap " circuit, shown in Fig. 6.8. V_1 and V_2 are initially cut off by negative grid bias. An input pulse generated at low power in a preceding circuit is fed to the grid of V_1, causing this valve to conduct. The current in R_1, which is a fairly high resistance, raises the voltage of the whole of the input circuit, and with it the grid of V_2, which thereupon conducts and allows C to discharge through the magnetron.

On the cessation of the pulse, both valves are again cut off. The diode V_3 allows C to charge during the resting period, and also damps out any reverse voltage swing across the magnetron at the end of the pulse.

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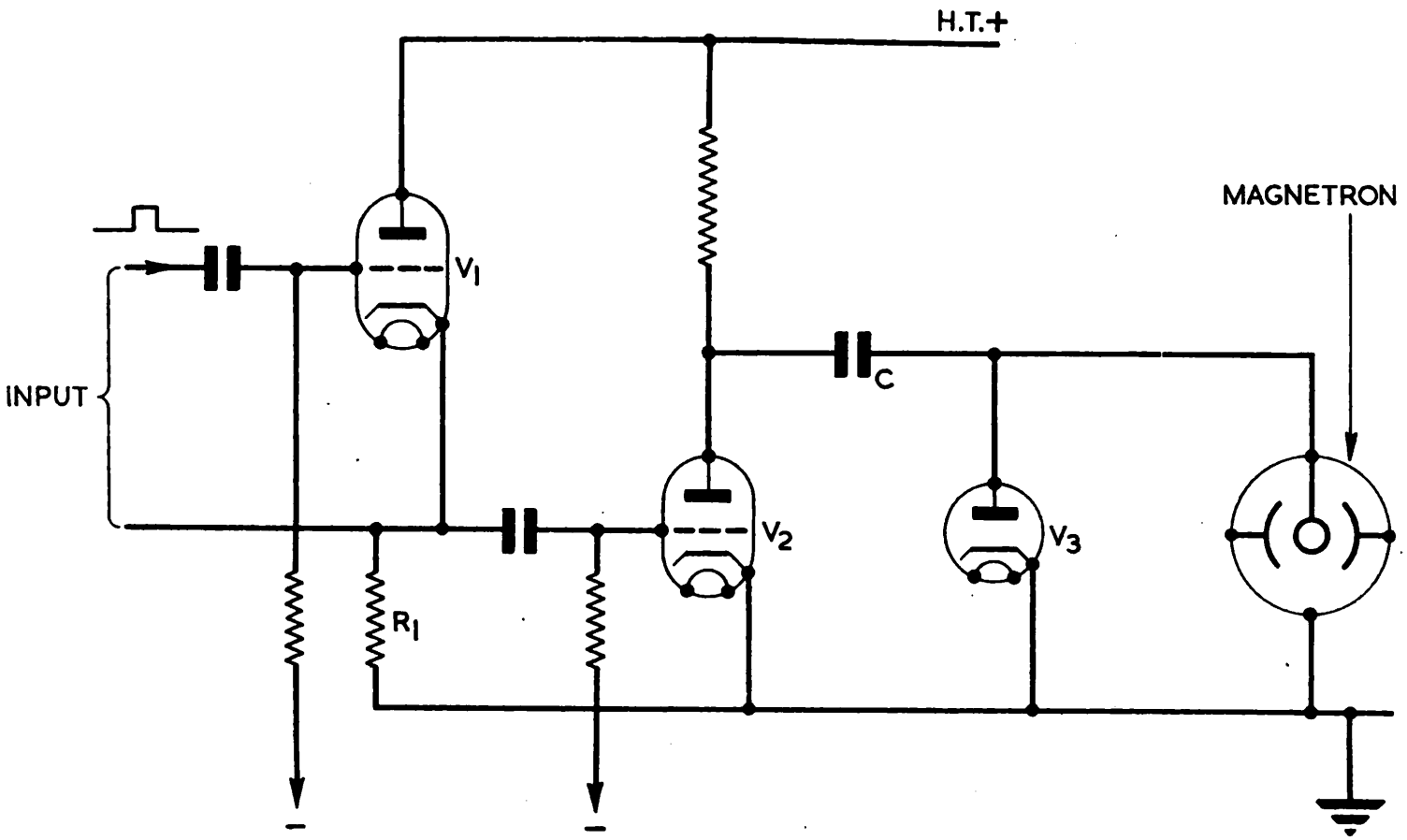

FIG. 6.8.—BOOTSTRAP CIRCUIT.

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The Magnetic Modulator

The magnetic modulator is based on the principle that the impedance of an iron-cored choke is high when the core is unsaturated and low when saturated. By arranging a circuit so that the choke passes from the unsaturated to the saturated

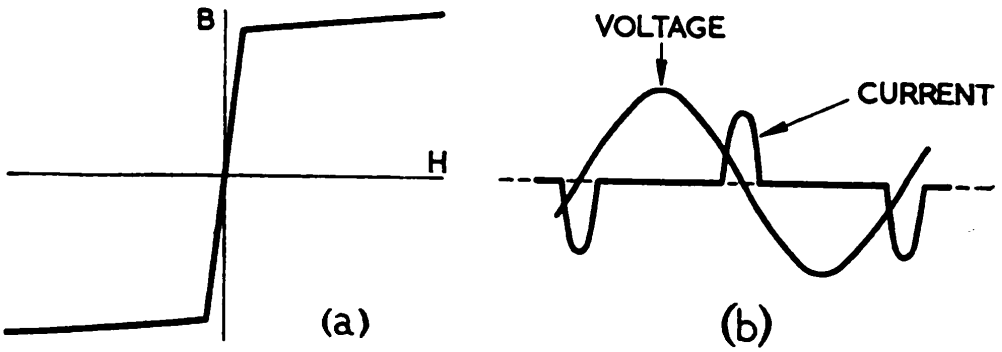

FIG. 6.9.—SATURABLE REACTOR.
(*a*) *B–H* curve.
(*b*) Voltage and current waveforms.

condition at the required moment, the effect of a switch is obtained. The "switch" is not, of course, a perfect one, because the unsaturated and saturated impedances are not infinite and zero respectively.

Magnetic materials of the mumetal type have very high permeabilities for flux densities up to a certain point, when the material saturates suddenly, and the incremental permeability

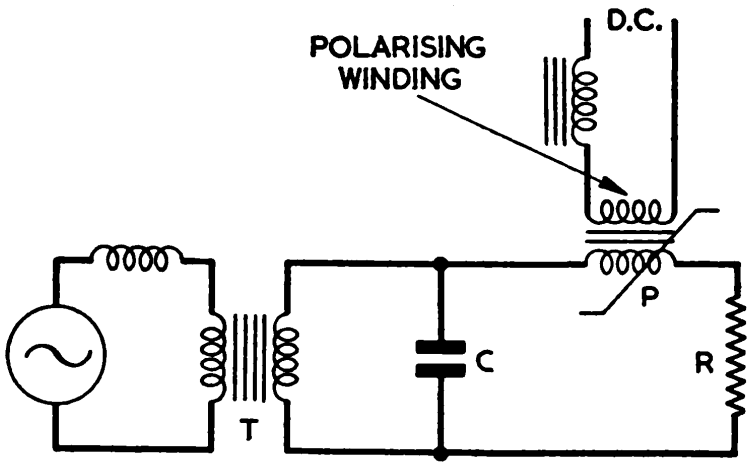

FIG. 6.10.—MAGNETIC MODULATOR—BASIC CIRCUIT.

becomes practically the same as that of air. The effective magnetization curve of such a material is shown in Fig. 6.9 (*a*).

An inductor having a core of this material, and supplied with a sinusoidal alternating voltage, will pass a current of the form shown in Fig. 6.9 (*b*). The sudden change of impedance when

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saturation occurs enables the device, which has been called a "pulsactor", to act as a switch.

Fig. 6.10 shows a circuit arranged for resonant charging of a capacitor C. A pulsactor P and load R in series are connected across C. The unsaturated impedance of the pulsactor (which is magnetically biased by a separate winding to ensure that saturation only occurs in one direction, i.e. only once per cycle of the supply frequency) is very high and has negligible effect on the voltage across C until the current in P has built up sufficiently to cause saturation. When this occurs P becomes

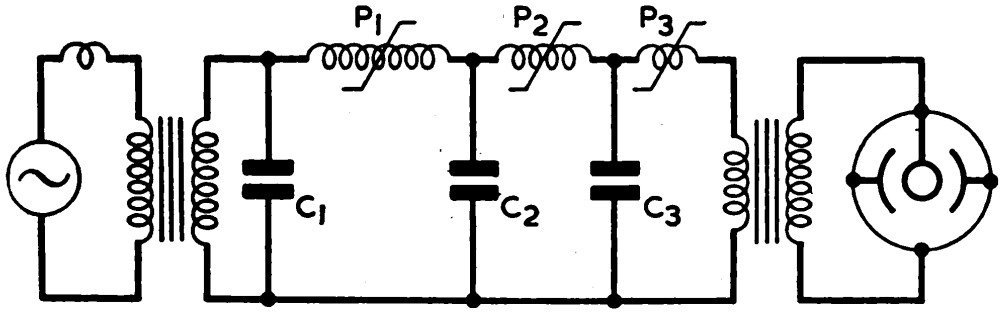

FIG. 6.11.—THREE-STAGE MAGNETIC MODULATOR.

a very small impedance and C discharges rapidly through the load R, after which the cycle is repeated.

The rate of rise and fall of the current in R, although rapid, is not normally high enough for producing a suitable pulse, and in practice it is necessary to replace R by further pulsactor stages, as shown in Fig. 6.11. In the last stage C_3 can be replaced by a suitable pulse-forming network, e.g. an artificial line, and the final load is the magnetron, fed through a pulse transformer. The rate of rise and fall of current in the pulsactors increases progressively from the input to the output end. Charging currents in the pulsactor windings usually render biasing from an external source unnecessary in all pulsactors except the first.

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CHAPTER VII

RADAR RECEIVERS

ALTHOUGH superregenerative receivers have been used for special purposes, the usual radar receiver is a superheterodyne. For metre wavelengths one or two stages of signal frequency amplification may be used, followed by a diode or crystal detector. For centimetre waves signal frequency amplification produces no useful increase in signal to noise ratio, and is not used, the signal being fed directly to a crystal mixer. This is followed by from six to ten stages of I.F. amplification, and the signal is finally amplified to the required level at video frequency.

To avoid swamping the input signal by interference which might be picked up by long lines, it is usual to place the first two or three stages of I.F. amplification (the head amplifier) as near to the mixer crystal as possible.

Receiver Sensitivity

The lower limit of useful signal input to the receiver is set by the " signal to noise " (S/N) ratio at the output. Since practically all the noise in the output originates at the input or in the first stage of the receiver, the addition of extra stages to increase the gain is only useful in so far as it raises the output to the required level. The signal/noise ratio, and therefore the minimum signal detectable, are not improved by this means.

Apart from the effects of interference from other electrical circuits, the noise in the output of a receiver is due to (*a*) noise coming in on the aerial from space, (*b*) thermal agitation noise, (*c*) shot effect in valves.

Thermal Agitation Noise

This is due to the random movements of electrons in a conductor, and its magnitude for a pure resistance is given by $\bar{E}^2 = 4kTR(f_2 - f_1)$, where $\bar{E}^2$ = the mean square voltage due to this random motion, k is Boltzmann's constant ($1{\cdot}38 \times 10^{-23}$ joules per degree), R is the resistance, T the absolute temperature in degrees Kelvin (273° K. = 0° C.), and f_1 and f_2 are the limits (in cycles per second) of the frequency band considered. Thermal agitation noise power is uniformly distributed over the entire frequency spectrum from zero to at least frequencies well above those concerning radar. For example, for a given pure resistance the noise power due to components whose frequencies lie between 0 and 1,000 c/s is the same as for those lying between 1 and 1·001 Mc/s. Noise power is proportional to the absolute temperature.

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For circuit elements which are not pure resistances, e.g. carbon composition resistors, crystals, etc., the noise power is higher than is given by the above formula by a factor which is called the " noise temperature ratio ". This leads to the conception of an " equivalent noise temperature " ; a device giving ten times as much noise as a uniform pure resistor of equal resistance is sometimes referred to as having an equivalent noise temperature of 2,900° K., assuming its actual temperature is 290° K., the figure usually taken as standard room temperature (because it gives a simple figure, viz. 4×10^{-21} joule, for the product of k and T).

Shot Effect

This is the noise produced by the random emission of electrons from a valve cathode. It is considerably reduced or " smoothed " by space charge effects. In tetrodes and pentodes there is a further random variation in anode current due to continuous minute variations in the proportions of electrons going to the anode and to the other electrodes. These valves are consequently more noisy than triodes.

Noise Factor of Receiver

The noise factor of a receiver may be defined briefly as the ratio of the noise/signal ratio at the output to that at the input, or, roughly, the ratio (noise gain)/(signal gain). The noise considered is that containing the frequency components lying within the bandwidth of the receiver ; the noise factor is consequently higher for greater bandwidth.

For an actual receiver the noise factor depends on circuit details, but may be of the order of 10 to 20, of which rather more than half is probably due to the amplifier and the remainder to the crystal mixer. Noise received from space on the aerial is approximately equal to that which would be produced by a pure resistance equal to the radiation resistance of the aerial ; compared with the noise produced by the crystal mixer, it is usually unimportant.

Crystal Mixers

A silicon crystal is generally used, and is accommodated in the transmit-receive portion of the aerial feeder system.

The local oscillator is a klystron, a fraction (about 1 milliwatt) of whose output is fed to the crystal, where it is mixed with the signal from the aerial. The conversion loss in the crystal is about 5 to 10 decibels, and its effective impedance about 300 to 600 ohms. The signal/noise ratio of a crystal increases as the standing current (due to the local oscillator) is increased from a low value, but it may reach a maximum before the maximum permissible current for the crystal is reached. The best adjustment in practice is found by trial.

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I.F. Amplifier

This may consist of about six to ten stages, having an overall gain of 80 to 100 decibels, the output being from a fraction of a volt to perhaps 10 or 20 volts, depending on the amount of video amplification provided. The input to the I.F. amplifier is of the order of microvolts.

Interstage coupling is by tuned inductances or transformers, the tuning being staggered to give the desired bandwidth. For example, if the I.F. is 45 Mc/s and a bandwidth of 10 Mc/s is required, the input and output may be tuned to 45 Mc/s and the interstage circuits to 40 and 50 Mc/s alternately.

Increasing the bandwidth improves the pulse shape, but results in increased noise output; in practice a compromise must be made between fidelity of reproduction of pulse shape and the attainment of a reasonable signal to noise ratio. A bandwidth, to "half power points", of $2/T$, where T is the pulse length, is generally satisfactory.

The actual value of the I.F. depends on the carrier frequency to some extent, intermediate frequencies of from 2 to 10 Mc/s being suitable for metre wavelengths, whilst 30 to 60 Mc/s is more appropriate for centimetric waves.

A diode or a germanium crystal may be used as second detector. Manual gain control can be effected by any of the usual methods, such as varying the grid bias on one or more stages, or varying the screen voltages.

Video Amplifiers

These are required to give a signal of sufficient intensity for modulating a cathode-ray tube (either by intensity modulation or deflection modulation, according to the type of display used), or to operate some auto-following or other special circuit.

In addition to receiving echo signals from the I.F. amplifier, the video amplifier will in many cases have to deal with calibration pips. In the case of P.P.I. presentation on a ship, a "heading line" will also be required to be shown. Most of the main points involved can be seen by considering a typical circuit (Fig. 7.1) of a video amplifier designed for P.P.I. presentation.

Negative going echo signals from the I.F. amplifier are amplified by V_1 and V_2, and passed, as negative pulses, to the cathode of the cathode-ray tube, causing momentary brightening of the trace. The inductances in the anode circuits provide some measure of high-frequency compensation to assist in preserving the pulse shape, and the crystal CR_2 limits the signal amplitude to an amount determined by the setting of R_5. R_4 limits the current passed by CR_2 when it conducts, thus avoiding spoiling the pulse shape by overloading V_1. R_4 is shunted by a small capacitance C_4 to pass the steep leading edge of the pulse. The crystal CR_1 allows C_3 to discharge rapidly on the cessation of the signal. The diode V_3 provides d.c. restoration for the

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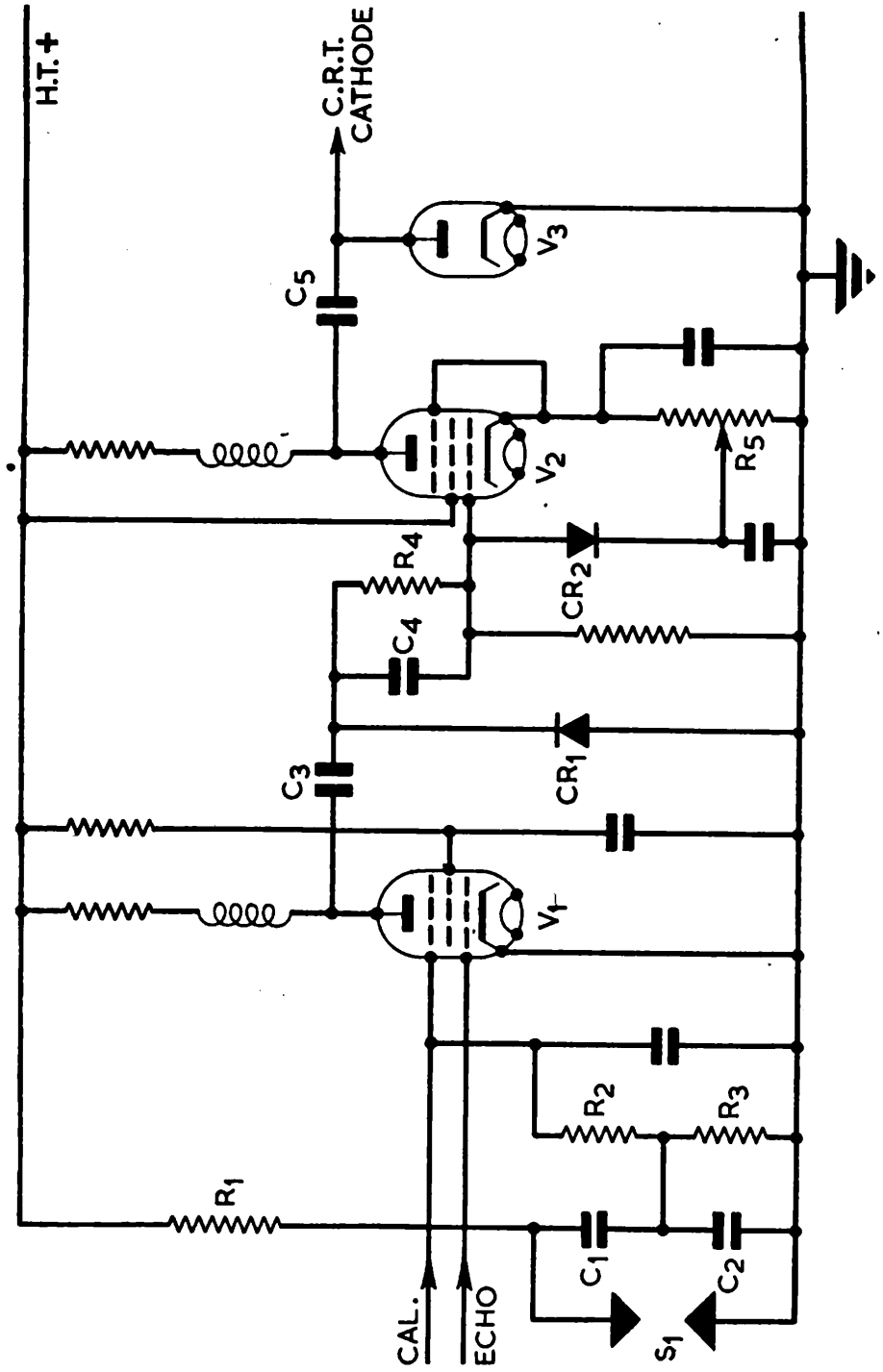

FIG. 7.1.—TYPICAL VIDEO AMPLIFIER.

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cathode of the cathode-ray tube, preventing it from going too positive when the echo signal ceases.

Calibration pips (negative going) are fed to the suppressor grid of V_1, cutting off this valve when they occur. The circuit $R_1C_1C_2R_2R_3$ provides for the introduction of a heading marker, produced by the closing of the switch contacts S_1 by the aerial when it is pointing dead ahead. This allows C_1 to discharge and produce a heading line on the screen.

There are many possible variations in a video amplifier circuit, depending on the nature of the installation and what it is required to do. The example given is merely one case, to show the principles involved.

Automatic Frequency Control (A.F.C.)

This is employed on most radars because it is important to maintain the intermediate frequency at its correct value, notwithstanding possible variations in the frequencies of klystron and magnetron.

Apart from drift due to temperature rise, the frequency of a magnetron can be appreciably affected by changes of load. These may occur as a result of powerful echoes from nearby objects being reflected back into the magnetron, also imperfections in rotating joints will produce variations in load. Other factors may also contribute their effects.

A satisfactory A.F.C. makes it practicable to keep the bandwidth of the receiver as small as possible, consistent with the preservation of pulse shape.

There are several possible A.F.C. circuits, and one example is shown in Fig. 7.2. The essential part of this circuit is a tuned discriminator transformer L_1L_2 tuned to the correct I.F. by C_1 and C_2.

The signal from the A.F.C. crystal mixer is suitably amplified and fed to the primary of the discriminator transformer, and also, through C_3, to the mid-point of the secondary. It can be seen from a.c. circuit theory that at resonance the voltage across L_2 is in quadrature with the input, and therefore the voltage applied to V_1 consists of two components, v_p and v_q respectively in phase and in quadrature with the input from the crystal (diagram *a*). The same applies to V_2 except that the quadrature component is in the opposite direction. The magnitudes of the voltages v_1 and v_2 across the two discriminator valves V_1 and V_2 are therefore equal, and the resulting rectified currents produce equal and opposite voltages across R_2 and R_3. The voltage on the grid of V_3 is therefore determined by the bias derived from the potential dividing circuit R_{11}, R_{12} through R_4. (Neglecting for the moment the effect of the *R–C* network $R_{5-7}C_{8-10}$.) The voltage on the repeller electrode of the klystron is determined by the current flowing in V_3 and the setting of R_{13}.

Now suppose that, for some reason, the I.F. falls below its

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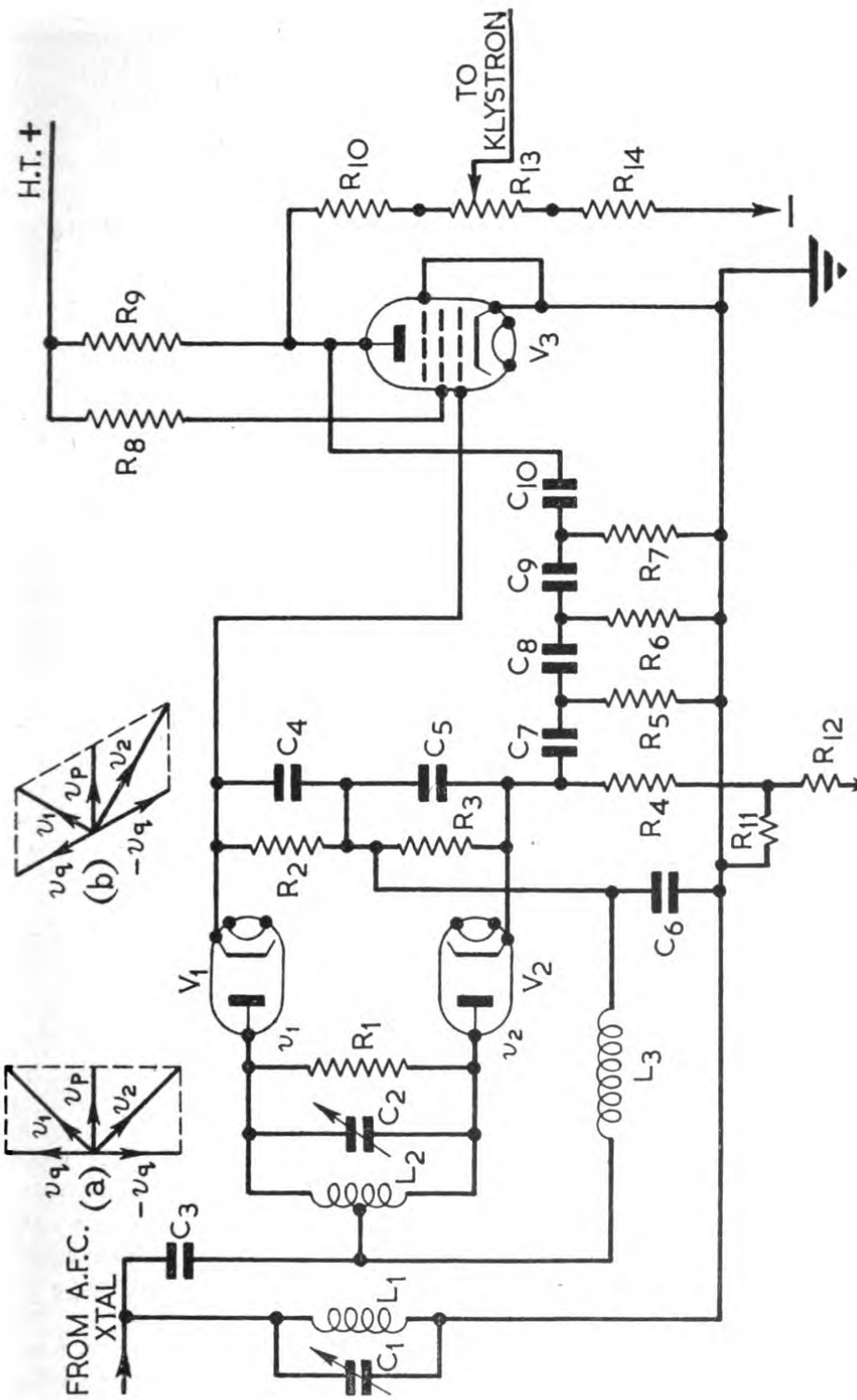

FIG. 7.2.—EXAMPLE OF A.F.C. CIRCUIT.

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proper value. The vector diagram (*b*) shows that the voltage across R_3 will now exceed that across R_2 and the voltage on the grid of V_3 will fall, reducing the anode current and so making the klystron repeller less negative, causing the frequency to fall, and so correct the error in the I.F. (It is assumed that this circuit applies to the case where the local oscillator frequency is below that of the magnetron. For the other case, either the diodes or the connections to the secondary of L_2 must be reversed.) The impedances L_3 and C_6 are respectively high and low at the intermediate frequency, and conversely at the pulse repetition frequency.

At frequencies considerably different from the correct I.F. the output of the discriminator circuit is negligible, and the circuit will not function. To deal with this, the *R–C* network $C_{8\text{-}10}R_{5\text{-}7}$ allows V_3 to act as an *R–C* phase shift oscillator. When oscillating (at the order of one cycle per second), this

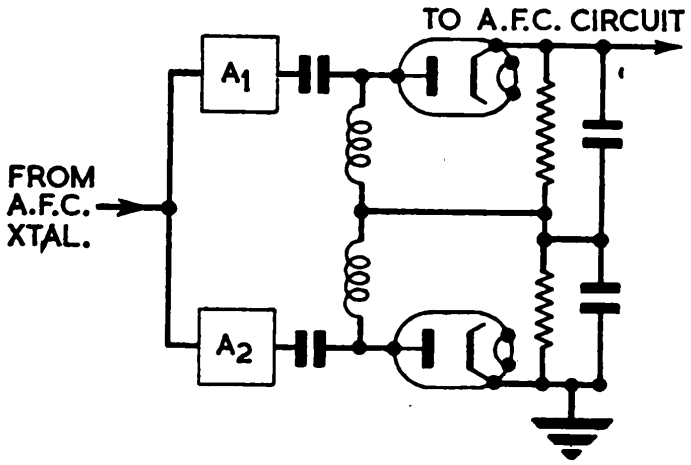

FIG. 7.3.—ALTERNATIVE TYPE OF DISCRIMINATOR FOR A.F.C. CIRCUIT.

sweeps the voltage of the klystron repeller through a sufficient range to ensure passing through the correct value for operation. When this occurs, and the grid of V_3 is going negative at the same time as the discriminator output is trying to drive it positive (or vice versa), the oscillation will be stopped and V_3 will thereafter function as a d.c. amplifier, and hold the klystron at its proper frequency by aid of the pulses from the discriminator as first described. The purpose of this oscillator is therefore to sweep the frequency over a fairly wide range to enable the A.F.C. to lock on, both initially and at any other time should the A.F.C. for any reason lose control. The circuit is initially set up by adjusting R_{13} with the discriminator and feedback circuits switched off.

In some circuits the particular type of discriminator transformer shown in Fig. 7.2 is replaced by two tuned amplifiers, one feeding V_1 and the other V_2 (see Fig. 7.3). These amplifiers are tuned to suitable frequencies, respectively above and below the correct frequency. At the correct I.F. the outputs are

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equal and oppose one another, whilst if the frequency deviates from the correct value, one amplifier produces a greater output than the other, and the difference controls the potential on the grid of V_3 as in the case previously described. Apart from this difference, the circuit functions in the same way.

A.F.C. Using Gas Tetrodes

Instead of the d.c. amplifier, R–C oscillator arrangement described above, a circuit using gas tetrodes (thyratrons) may be employed.

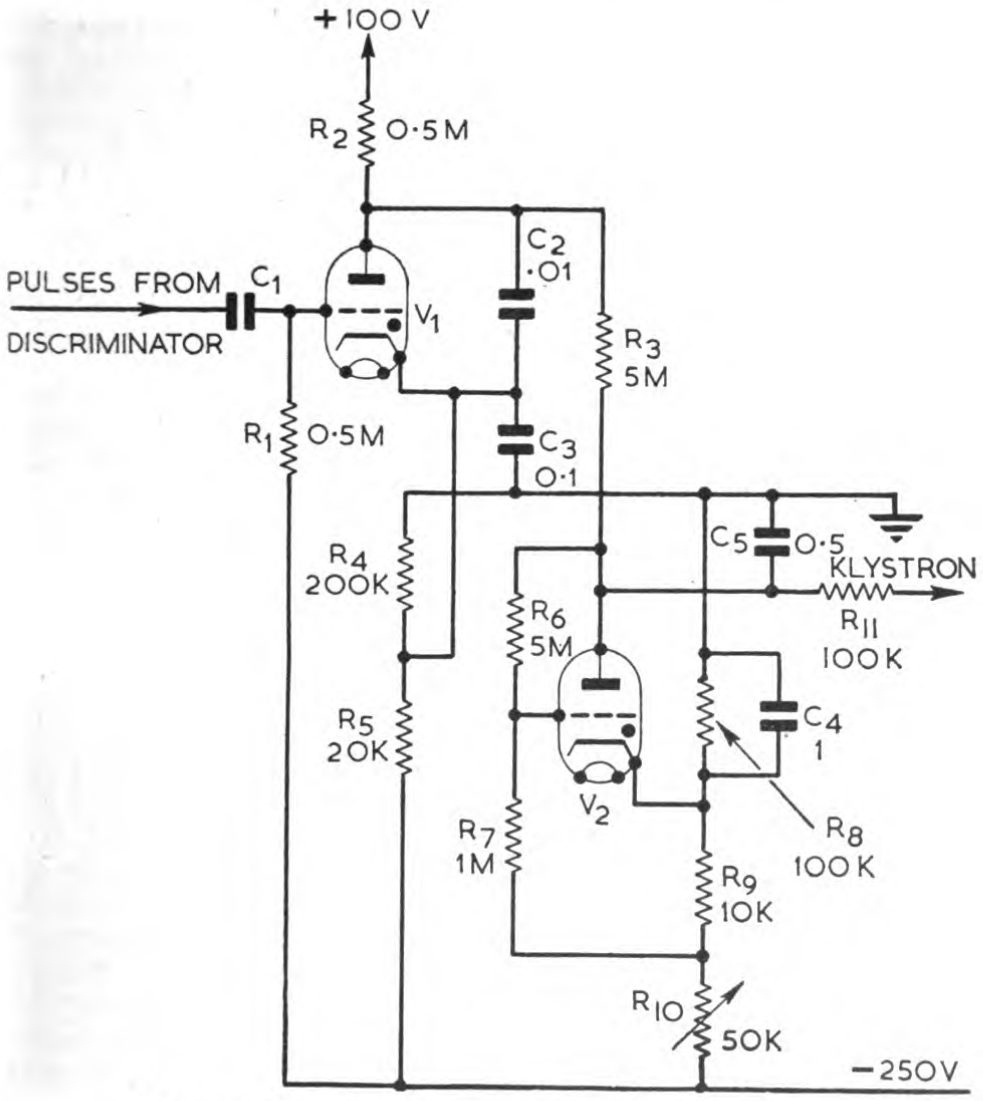

FIG. 7.4.—A.F.C. CIRCUIT USING GAS TRIODES.

The output from a discriminator of either of the types described is amplified (at video frequency) and fed, in the form of pulses at the pulse repetition frequency, to the circuit shown in Fig. 7.4. The connections must be such that the pulses are

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positive when the klystron frequency is too low, and negative when it is too high, and of course zero at the correct frequency.

In the absence of pulses, V_1 is cut off by negative bias on its grid. C_5 charges through the resistance chain $R_2R_3R_6R_7$, and when its voltage has risen sufficiently, V_2 conducts, its bias having a suitable value, derived from $R_8R_9R_{10}$ to permit this. The voltage on C_5, and therefore on the repeller of the klystron, falls until V_2 ceases to conduct, and the cycle is repeated. The voltage applied to the klystron is thus swept over a fairly large range at a relatively low frequency (say about once per second). At some time during this sweep, when the klystron voltage is rising and its frequency is falling, positive pulses will be received from the discriminator just after the klystron frequency has passed through its correct value in the downward direction. These pulses are of sufficient amplitude to make V_1 conduct, whereupon the resulting fall of potential at the positive end of the resistance chain R_3 . . . R_7 will arrest the rising voltage on C_5. This capacitor then starts to discharge through R_3 and the klystron frequency rises, and when it is slightly above the correct value, negative pulses are produced by the discriminator. These pulses cut off V_1, which they are able to do because by this time C_2 will have discharged and the anode current will be very small. The potential at the anode of V_1 therefore rises, and with it that of C_5, by charging through R_3, etc. This causes the klystron frequency to fall once more, and the cycle is repeated. The frequency is thus swept rapidly, say 50 to 100 times per second, over a small range, within the bandwidth of the receiver, on each side of the correct value. V_2 plays no further part unless for some reason V_1 loses control. The function of V_2 is simply to sweep the frequency over a wide range to bring it within the limits with which V_1 can deal.

The component values marked on the diagram are only intended to give an indication of the magnitudes involved. R_{10} is adjusted on setting up to ensure that the klystron oscillates in the desired mode.

Swept Gain (Anti-Clutter)

To avoid paralysis of the receiver by direct pick-up of the transmitter pulse, and excessive clutter on the display due to powerful echoes from nearby objects, it is necessary to cut off the receiver during the transmitter pulse, and to operate it at reduced gain for a few microseconds after the end of each pulse. This is done by using a " swept gain " circuit.

Swept gain circuits differ in detail and complexity with different radars, but that shown in Fig. 7.5 illustrates the basic principle.

In the quiescent state, V_2 is conducting and its anode, and therefore the bias supply to the I.F. stages is at substantially earth potential and the gain of the receiver is normal.

When the transmitter fires a negative trigger pulse, of the

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order of 20 to 100 volts, is applied to the cathode of V_1. This pulse can be obtained from an auxiliary winding on the magnetron pulse transformer, or may be tapped off across a resistor in the earth connection of the magnetron. The pulse causes V_1 and V_3 to conduct, but as the pulse is large the drop across these valves will be sufficient to make the bias negative enough to cut off the receiver. On the cessation of the pulse, the bias voltage rapidly rises (i.e. becomes less negative) until V_3 ceases to conduct. The voltage at which this occurs is determined by

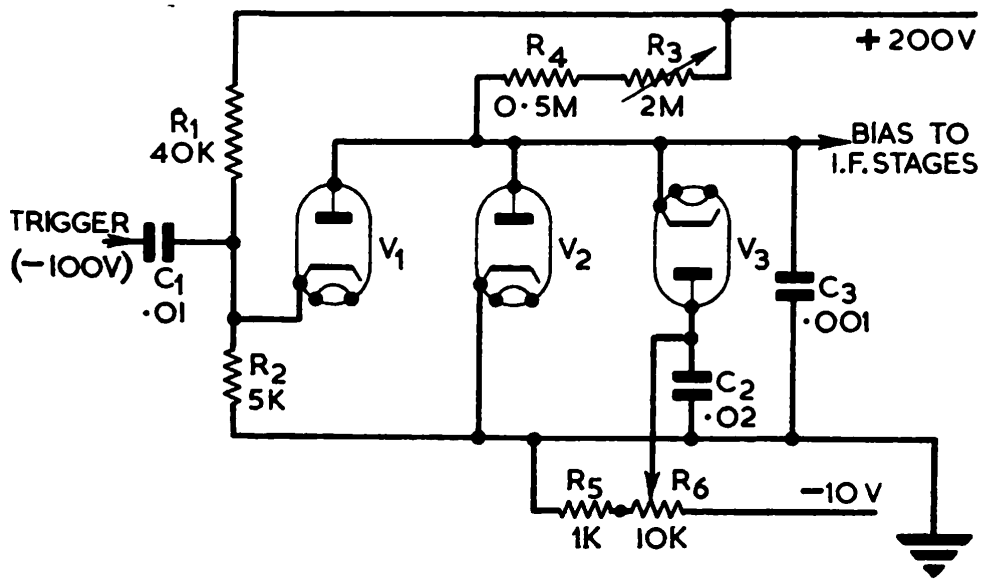

FIG. 7.5.—SWEPT GAIN CIRCUIT.

the setting of R_6. Thereafter the voltage continues to rise at a slower rate, determined by the time C_3 takes to charge through R_3 and R_4. The rise continues until V_2 conducts, when the bias is zero and normal gain is restored. In the case shown this may take anything from approximately 12 to 125 microseconds, depending on the settings of R_3 and R_6. It should be noted that the rate of rise is controlled by R_3, but that the point from which it starts depends on R_6.

The component values given in the diagram are only intended to indicate approximate magnitudes.

Differentiation (Anti-Clutter)

In some cases, particularly in marine radar with P.P.I. presentation, echoes from large land masses at close range, from rough seas, or from heavy rain may produce the appearance of masses of "solid clutter" on the screen. This can cause small objects to be obscured, and the picture made generally less intelligible.

In these circumstances a simple switching operation which reduces the otherwise relatively large time constant of one of the video interstage couplings to a small value, of the order of

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the pulse length (thus introducing differentiation), will break up solid masses of clutter and result in considerable cleaning up of the picture. The gain is somewhat reduced, but this can be compensated for by switching in a different value of resistance at some convenient point in the circuit.

Automatic Gain Control

Since the average output of a radar receiver consists mainly of noise, A.G.C. as used in communication work would operate almost entirely on noise, and would not be of much value. A.G.C. circuits have, however, been fitted in some radars for the purpose of reducing gain when large masses of clutter are present. The result is somewhat similar to that obtained with differentiation.

A different case is the application of A.G.C. to one particular echo signal which is being followed. (See Chapter XV, page 155.)

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CHAPTER VIII

TRANSMISSION LINES (FEEDERS)

In radar these are required for conveying the R.F. energy from the transmitter to the aerial and from the aerial to the receiver.

There are three main types: the open-wire feeder, the concentric or coaxial feeder, and the waveguide.

The first type, although used in the past for certain metre wave equipments, is now practically obsolete, largely because at very short wavelengths the losses by direct radiation from the feeder are excessive. Apart from actual energy loss, considerations of interference would prohibit their use on modern radars.

Concentric feeders can be used for wavelengths ranging from relatively large values down to about 10 centimetres, after which the waveguide is generally used. There is no sharp transition point, and for wavelengths round about 10 centimetres either may be used.

Propagation along Uniform Line

If an alternating voltage is applied to the end of an infinitely long uniform line, the voltage at any point along the line will vary sinusoidally with time, and there will be a phase difference

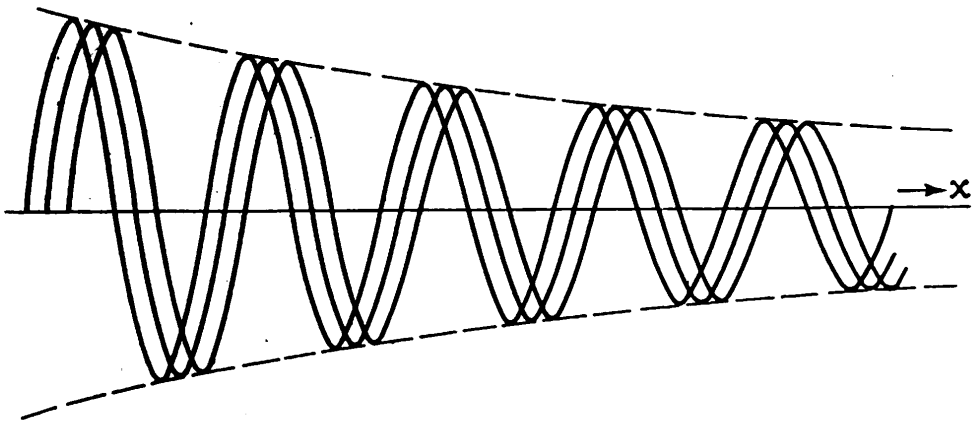

Fig. 8.1.—Propagation of Voltage Wave along Uniform Line.

between the voltages at different points at any one instant. If there is attenuation, there will also be a difference of amplitude.

At any one instant, the instantaneous voltage will vary sinusoidally with distance along the line, subject to a decrease in amplitude due to attenuation.

These points are illustrated in Fig. 8.1, which shows the voltage

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distribution along a uniform line for three different instants, at one-eighth cycle intervals. The pattern takes the form of a wave moving with uniform velocity along the line. Similar remarks apply to the current in the line.

If the voltage applied to the input end is $V \sin \omega t$, the voltage at a point distant x along the line is $Ve^{-\alpha x} \sin (\omega t - \beta x)$. α is called the attenuation constant and β the phase constant. The complex quantity $P = \alpha + j\beta$ is called the propagation constant.

The above expression for voltage shows that the wave travels along the line with velocity ω/β.

The input impedance of an infinite length of line is finite, and is usually denoted by Z_0, called the characteristic impedance. The current in the line at any point is equal to the voltage at that point divided by Z_0. (*N.B.* If the line is of finite length, but is terminated by a load of impedance equal to Z_0, the input impedance is also Z_0.)

Line Constants

If L, C, R, G are respectively the series inductance, shunt capacitance, series resistance and shunt conductance (all per unit length of line), it can be shown that

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \qquad P = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}.$$

For feeders used in radio work, R and G are respectively small compared with ωL and ωC and the following simpler formulae can be used:

$$Z_0 = \sqrt{\frac{L}{C}} \qquad \alpha = \frac{1}{2}\left(\frac{R}{Z_0} + Z_0 G\right) \qquad \beta = \omega\sqrt{LC}.$$

It will be noted that Z_0 is a pure resistance, and that attenuation is *apparently* independent of frequency. This is not the case in practice because, as the frequency increases, R increases due to skin effect and G (representing dielectric losses rather than ohmic leakance) also increases.

Impedance Mismatch—Reflection

If the line is of finite length and is terminated by a load of impedance Z_r (not equal to Z_0), the voltage and current waves will be partially reflected at the load, giving rise to waves travelling back towards the sending end. These reflected waves can be represented by formulae similar to those for the forward waves, except that the term $\sin (\omega t - \beta x)$ must be replaced by $\sin (\omega t + \beta x)$ since the wave is travelling in the opposite direction. Similarly $e^{-\alpha x}$ is replaced by $e^{\alpha x}$.

The reason why reflection occurs can be explained as follows. Let V_1 and V_2 be the amplitudes of the forward and reflected

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waves, at the point where the load is connected, and let I_1 and I_2 be the corresponding currents.

From the fundamental fact that, in a uniform line, the amplitude of the voltage wave is Z_0 times that of the current wave, we have

$$V_1/I_1 = Z_0 \qquad \text{and} \qquad V_2/I_2 = Z_0 .$$

But the total voltage across the load is $V_1 + V_2$ and the current in the load is $I_1 - I_2$,

whence $$(V_1 + V_2) = Z_R(I_1 - I_2).$$

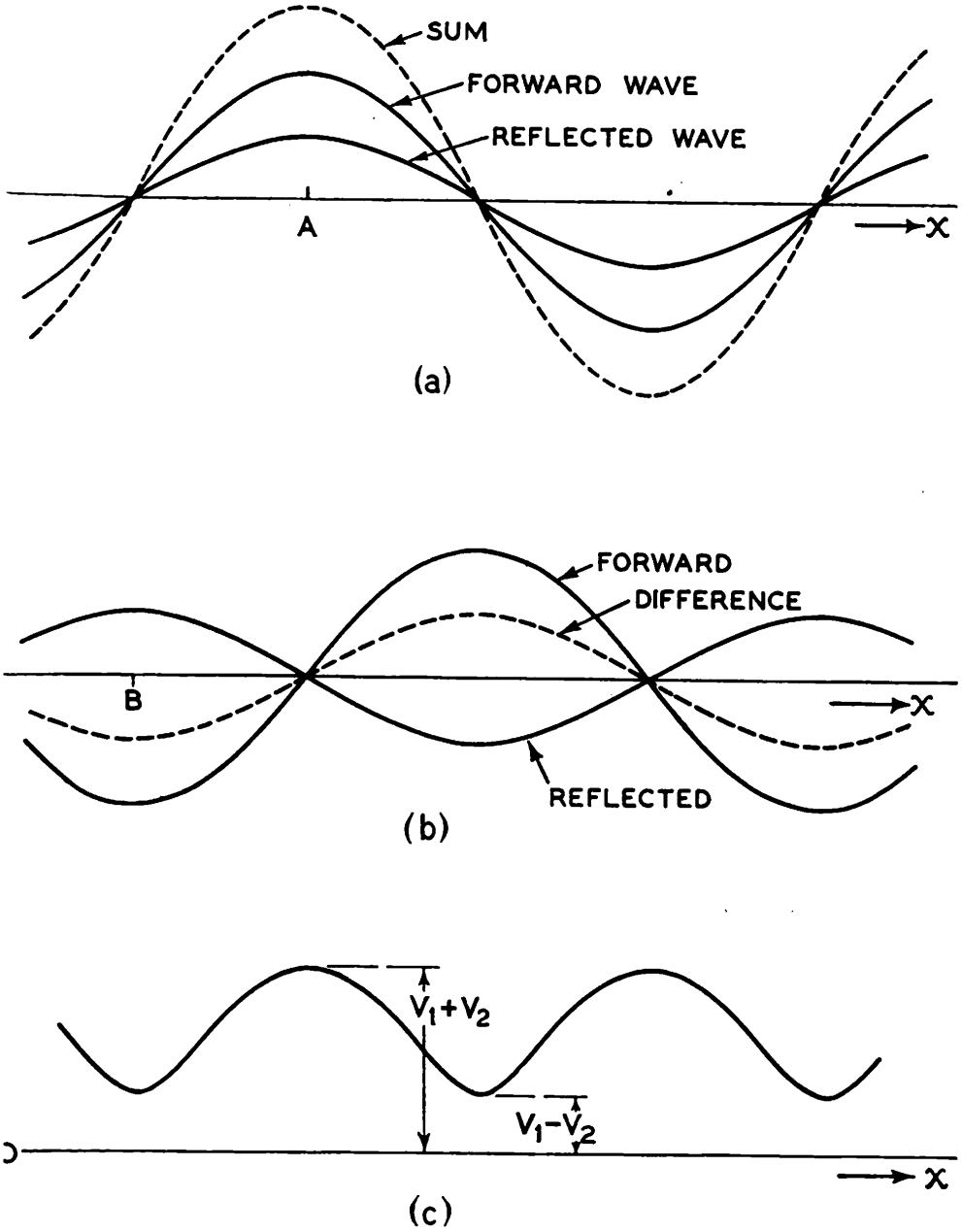

FIG. 8.2.—STANDING WAVES CAUSED BY INTERFERENCE BETWEEN FORWARD AND REFLECTED WAVES IN UNIFORM LINE.

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Solving these equations, it is found that

$$V_2/V_1 = \frac{Z_r - Z_0}{Z_r + Z_0} = K \text{ (the reflection coefficient).}$$

If Z_r is a pure resistance, K is a real number, positive or negative according as Z_r is greater or less than Z_0. A positive value means that the voltage wave is reflected without change of phase, a negative value implies reversal (180°) of phase on reflection.

If Z_r is complex, equal to $(R_r + jX_r)$, say, K is also complex, and its magnitude is

$$k = \left[\frac{(Z_0 - R_r)^2 + X_r^2}{(Z_0 + R_r)^2 + X_r^2}\right]^{\frac{1}{2}}.$$

If there is any discontinuity of impedance, such as a load connected across a line at some intermediate point, the forward wave will be partially reflected at the discontinuity and partially transmitted to the line beyond.

Standing Waves

Forward and reflected waves will "interfere", and a series of points, one half-wavelength apart, can be found at which both waves reach their maxima in the same direction at the same instant. At other points, midway between these, the waves will reach their maxima in opposite directions at a time one-quarter of a cycle later. These conditions are shown in Fig. 8.2 (*a*) and (*b*) respectively. The maximum voltage at point A is the sum of the maxima of forward and reflected waves, whereas at B it is their difference. (*c*) shows the way in which the resultant maximum varies at different points along the line. The ratio of the heights $(V_1 + V_2)$ and $(V_1 - V_2)$ is the standing-wave ratio (S.W.R.). This is usually denoted by S and is equal to $(1 + k)/(1 - k)$.

Sometimes S is defined as $(1 - k)/(1 + k)$, which is the reciprocal of the former value. No confusion can result from this non-uniformity of practice because, in the first case S can never be less than unity, whereas in the second it can never be greater.

The current standing-wave ratio is numerically equal to the voltage S.W.R., but points of maximum current are points of minimum voltage, and vice versa.

Input Impedance

The input impedance of a line of length l terminated by a load Z_r is $Z_i = Z_0\left(\frac{Z_r \cosh Pl + Z_0 \sinh Pl}{Z_0 \cosh Pl + Z_r \sinh Pl}\right)$

where $\cosh Pl = \frac{e^{Pl} + e^{-Pl}}{2}$ $\sinh Pl = \frac{e^{Pl} - e^{-Pl}}{2}$.

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If R and G are neglected, the input impedance becomes

$$Z_i = Z_0\left(\frac{Z_r \cos \beta l + jZ_0 \sin \beta l}{Z_0 \cos \beta l + jZ_r \sin \beta l}\right).$$

The ratio $\frac{\text{voltage at input end of line}}{\text{current in load}}$, called the receiving-end impedance, is $Z_0 \sinh Pl + Z_r \cosh Pl$, which reduces to $Z_r \cos \beta l + jZ_0 \sin \beta l$ if R and G are neglected. If $Z_r = Z_0$ the input and receiving-end impedances are equal in magnitude.

In the special case where $l = \lambda/4$, $Z_i = Z_0^2/Z_r$, whence a quarter-wavelength of line can be used as an impedance transformer. If, in this case, the far end is open circuited, the input impedance is zero (equivalent to a series resonant circuit), whilst if the far end is short circuited, the input impedance is infinite (parallel resonance). In practice, owing to R and G not being zero, the impedances in the two cases will be very small and very large, instead of zero and infinite, respectively.

Close approximations to the actual values are $Z_0\alpha l$ and $Z_0/\alpha l$ respectively.

Relation between Load Impedance and S.W.R.

The impedance of a load can be determined, in terms of that of the feeder, by observing the magnitude of the S.W.R. and the positions of the voltage minima. The formula is

$$Z_r = Z_0\left[\frac{2S - j(S^2 - 1)\sin \phi}{(S^2 + 1) + (S^2 - 1)\cos \phi}\right]$$

where $\phi = \frac{4\pi a}{\lambda}$ radians $\left(720\frac{a}{\lambda} \text{ degrees}\right)$, and a is the distance of the first voltage minimum from the load.

The impedance at the input end of a feeder can be calculated from the formula

$$Z_0\left[\frac{2S + j(S^2 - 1)\sin \theta}{(S^2 + 1) + (S^2 - 1)\cos \theta}\right]$$

where $\theta = \frac{4\pi b}{\lambda}$ radians, and b is the distance from the input end to the nearest point of voltage minimum.

The corresponding formula in terms of k is

$$Z_0\frac{(1 - k^2) + j2k \sin \theta}{(1 + k^2) + 2k \cos \theta}.$$

The Smith Diagram

Many of the foregoing calculations can be carried out graphically by aid of the Smith diagram, shown in skeleton form in Fig. 8.3. The diagram consists of three sets of circles representing reflection coefficient k, resistance R/Z_0 and reactance X/Z_0 (the last two being expressed in terms of Z_0).

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The radii of the circles are respectively equal to

$$k, \qquad \frac{1}{1 + R/Z_0}, \qquad Z_0/X.$$

For a uniform line (Z_0) terminated by a load $Z_r = R + jX$, the impedance looking towards the load at any point on the line is represented by the appropriate point on the corresponding constant-k circle.

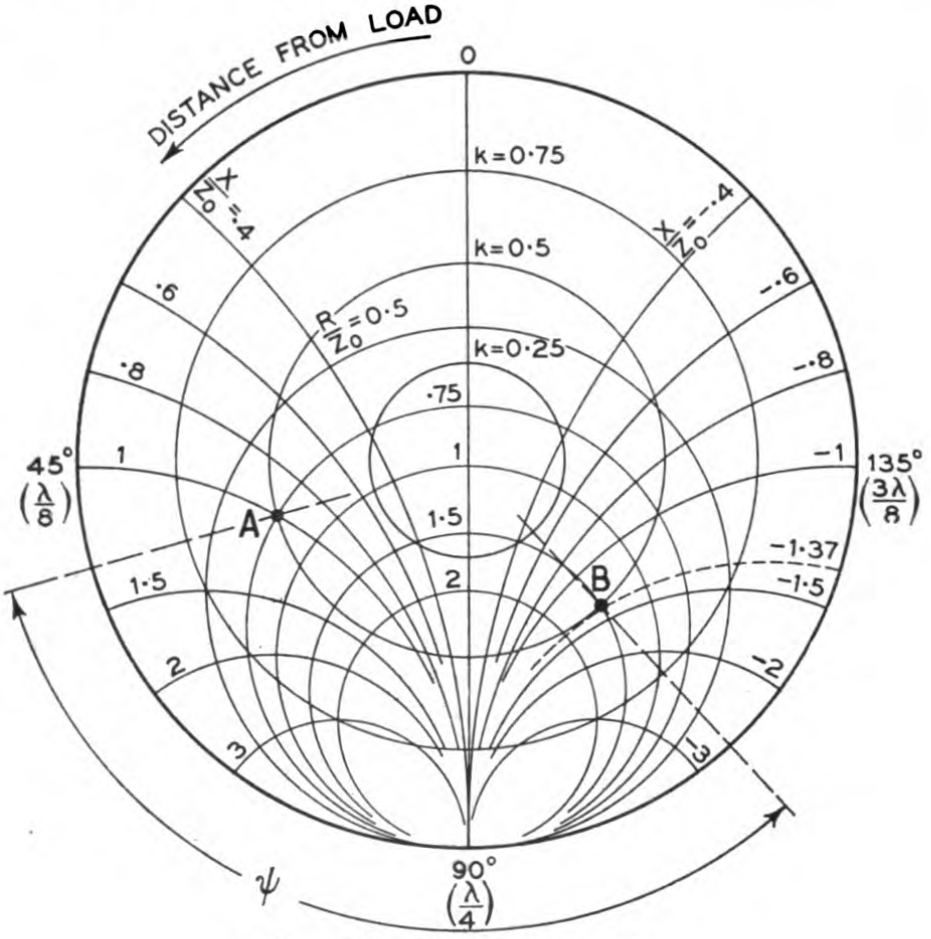

FIG. 8.3.—SMITH DIAGRAM.

Example.

Let A represent the load (in this case $R/Z_0 = 0{\cdot}75$, $X/Z_0 = 1$, $k = 0{\cdot}5$), the impedance at distance x away, measured away from the load, is represented by point B, and is $1{\cdot}5 - j1{\cdot}37$. (The angle $\psi = \frac{360k}{\lambda}$. It is the distance measured in electrical degrees.)

If the diagram is printed for general use, it contains as many circles as can conveniently be drawn in without causing confusion due to overcrowding; but if specially drawn for a particular case, it need only contain those circles required for the problem being solved.

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Note that one complete revolution round the diagram corresponds to 180 electrical degrees (π radians) or one half-wavelength.

Impedance Matching

Impedance mismatch is generally undesirable for a variety of reasons, such as the reflected power interfering with the working of an oscillator, the higher voltages resulting from standing waves increasing the liability to breakdown without any corresponding advantage, etc.

As a short-circuited length of line acts as a reactance of value depending on its length, two stubs placed near the load

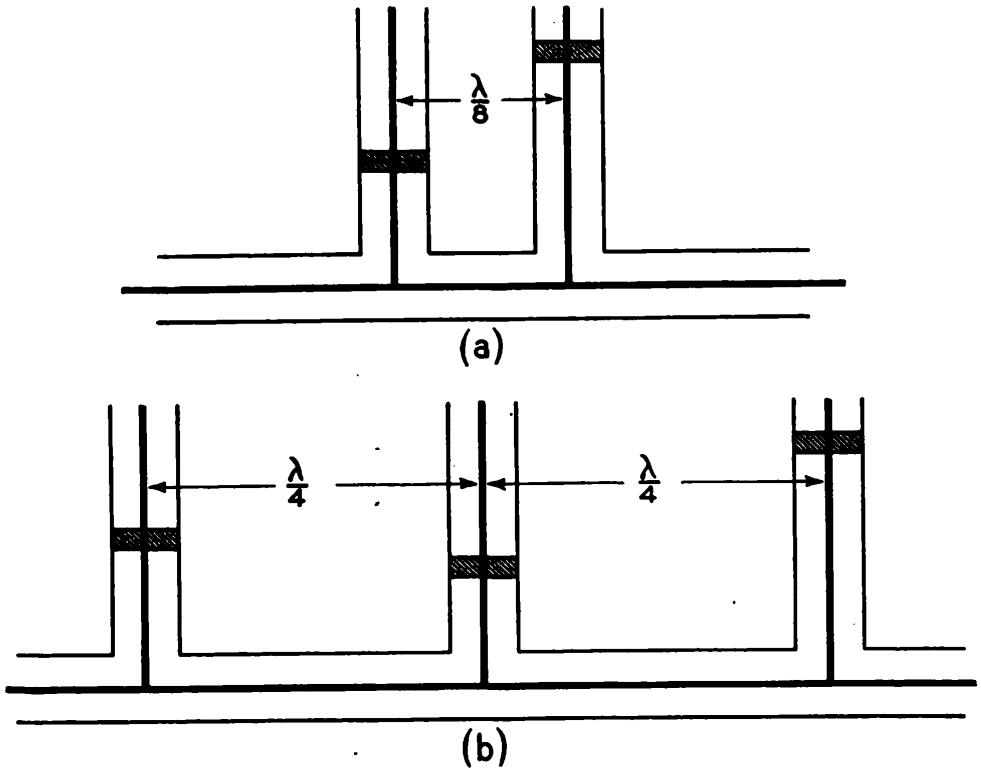

FIG. 8.4.—IMPEDANCE MATCHING USING ADJUSTABLE STUBS.

[Fig. 8.4 (*a*)] can be used for correcting mismatch. The lengths of the two stub lines are independently adjusted, by moving the short circuiting plungers until the required match, shown by S.W.R. = 1, is obtained. This arrangement permits matching only within certain limits, but that shown in Fig. 8.4 (*b*), using three stubs permits, theoretically, the correction of any degree of mismatch, although the practical difficulties are great if the mismatch is considerable.

Long Line Effect

The input impedance of a feeder line connected to a load not properly matched depends, among other things, on the electrical

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length of the line. A slight change of frequency will alter the electrical length and therefore the input impedance. This change of impedance may affect the frequency of an oscillator, e.g. a magnetron, connected to the input end, and instability may result if the feeder is long enough for the effect described to be appreciable.

Fig. 8.5 shows the variation of input impedance of a long feeder with frequency. The straight line I cutting the curve represents the relation between the frequency of an oscillator and the reactance connected to it, i.e. the input reactance of the feeder. This line I cuts the curve in three points, *A*, *B*, *C*. *B* is unstable, but *A* and *C* both represent possible stable conditions. There will thus be an uncertainty whether the oscillator will operate at the frequency corresponding to point *A* or *C*. Successive pulses may, in fact, be on different frequencies.

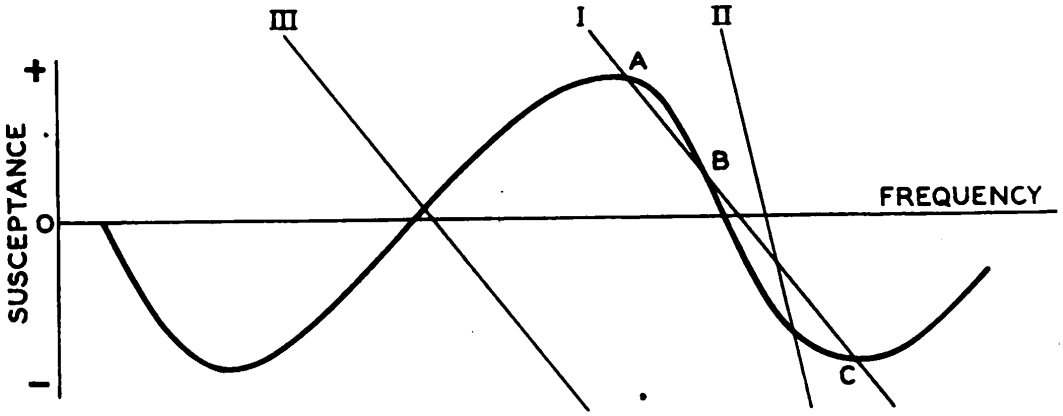

FIG. 8.5.—LONG LINE EFFECT.

The cure for this trouble is either to use an oscillator less affected by changes in the connected impedance or to shorten the feeder so that its input impedance is less affected by frequency changes over the working range. Either of these gives a line (II) which cuts the curve in one point only.

If neither of these expedients is acceptable, but the length of the feeder is altered slightly so that line I moves to the position occupied by line III, it will cut the curve in one point only, and satisfactory operation should be practicable; but in this case some phasing device must be incorporated in the feeder and adjusted to bring its electrical length to a suitable value every time the oscillator is changed.

Note that if the feeder is so long that the transmitter pulse has ceased before the reflected wave returns from the far end, these considerations do not arise.

Coaxial Feeders—Construction

If required to be flexible, a concentric cable with a suitable insulant such as polythene may be used.

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Rigid feeders take the form of an outer tube with the inner conductor supported on dielectric spacers.

Another method is to support the inner conductor on quarter-wave stubs, which have infinite impedance at resonance. By making the diameter of the inner conductor larger at the point of support, as shown in Fig. 8.6, some compensation is obtained for the fact that the stub has an infinite impedance at only one frequency. This expedient permits reasonable operation over a frequency band of total width equal to 20 to 30 per cent of the mid-frequency.

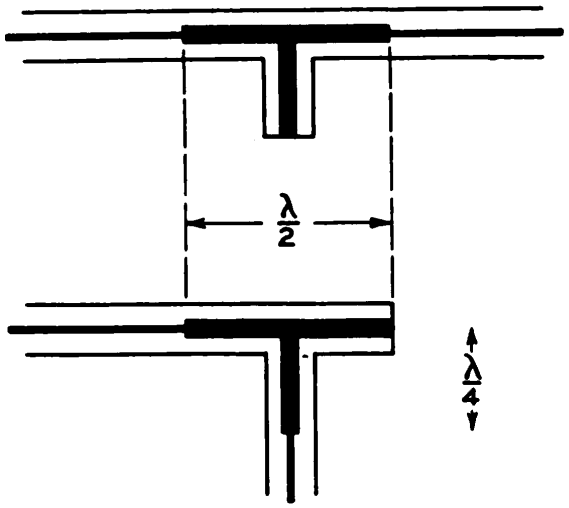

FIG. 8.6.—SUPPORTING STUBS FOR COAXIAL FEEDER.

Limitations on Dimensions

The impedance of a coaxial line is proportional to the logarithm of the ratio of the diameters of outer and inner conductors. For a given impedance this ratio is therefore fixed. For high powers, e.g. in transmitter feeders, the spacing must be great enough to prevent sparking over, a possibility aggravated by a poor S.W.R. The inside diameter of the outer conductor

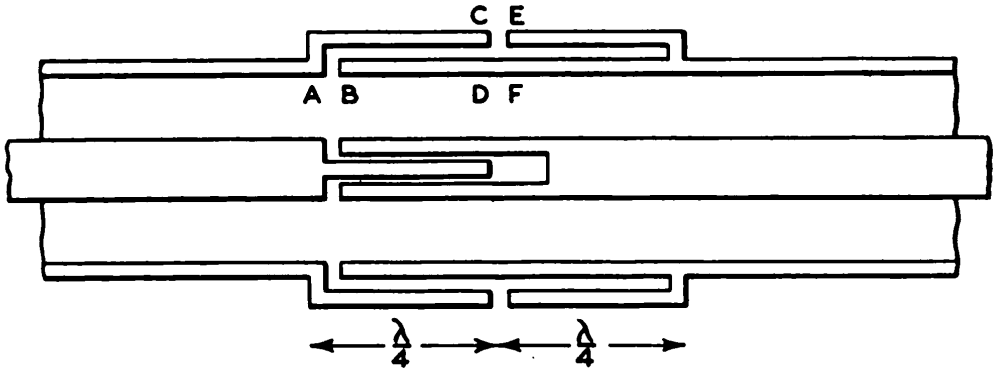

FIG. 8.7.—ROTATING JOINT FOR COAXIAL FEEDER.

should not however exceed approximately 0·6 times the working free-space wavelength or there will be danger of the line acting as a waveguide as well as a concentric line. Interference between the two modes of transmission would result in unacceptable conditions. This refers to air insulation. For solid insulation the figure is $0{\cdot}6/\sqrt{\epsilon}$, where ϵ is the permittivity of the insulant relative to air.

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Rotating Joints

In Fig. 8.7 the impedance between E and F is infinite because these two points are at the end of a quarter-wave line short circuited at its far end. The impedance between C and D is therefore also infinite, notwithstanding that there may be some accidental connection between C and E, due to dirt, etc., or even the deliberate addition of a dirt-excluding ring. The impedance between A and B is therefore zero and the effect of a perfect connection between the two halves of the joint is obtained without the use of rubbing contacts, which are liable to be unsatisfactory in practice.

Unbalance to Balance Converter

One conductor of a coaxial line is necessarily at earth potential as regards R.F. If a balanced feed, e.g. to a dipole, is required, the device shown in Fig. 8.8 can be used. It consists of the

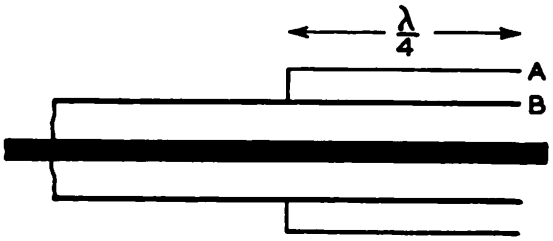

FIG. 8.8.—UNBALANCE TO BALANCE CONVERTER.

addition of a second outer conductor $\lambda/4$ long. Since the impedance between the points A and B is infinite, the original outer conductor is effectively isolated from earth at its output

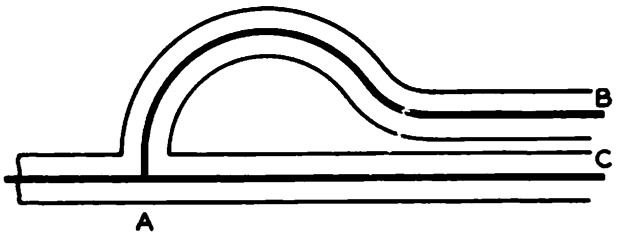

FIG. 8.9.—ALTERNATIVE FORM OF UNBALANCE TO BALANCE CONVERTER.

end. This device is sometimes called a balun, a bozooka, or a $\lambda/4$ can.

Another method is shown in Fig. 8.9, the lengths of the two sections A–B and A–C differing by half a wavelength.

Artificial Lines

A network of any of the types shown in Fig. 8.10 can be used to simulate a short length of uniform line, and a complete artificial line can be built up by adding a number of sections in series.

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The equivalence is not exact, but is reasonably good provided each section represents not more than a small fraction of a wavelength.

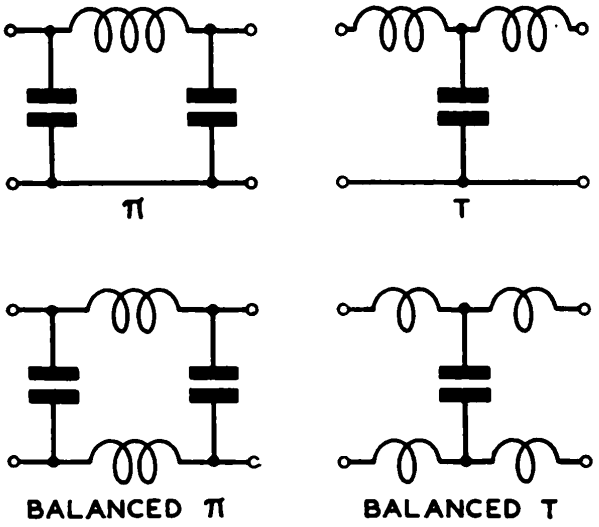

FIG. 8.10.—ARTIFICIAL LINE SECTIONS.

Artificial lines have no applications as feeders but are useful for introducing a delay into a circuit when the necessary length of actual line, to produce the required delay, would be inconveniently great. They are also used in pulse-forming circuits.

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CHAPTER IX

WAVEGUIDES, HORNS AND CAVITY RESONATORS

WAVEGUIDES are only practicable for the higher frequencies because their dimensions must be of the same order as the wavelength being used. For wavelengths around 10 centimetres either waveguides or coaxial lines may be used as feeders, but at shorter wavelengths, e.g. 3 centimetres, the waveguide is used almost exclusively because of its lower attenuation and greater power-handling capacity at very high frequencies.

Practical waveguides consist of metal tubes of rectangular or circular cross-section, the former being by far the commoner. To keep down attenuation, high conductivity copper is used, and in some cases the inner surface is silver plated to improve the conductivity still further.

Mode of Transmission

Radiation travels in a waveguide according to the same fundamental laws as in free space, but the field patterns are modified because, at the surface of a perfect conductor, there can be no tangential component of electric force nor normal component of magnetic force. (For the purpose of this statement copper may be regarded as a perfect conductor.) Furthermore, it can be shown that for propagation to be possible, there must be a component of either electric or magnetic force in the direction of propagation, unlike the case of radiation in free space where the electric and magnetic fields are wholly transverse to the direction of propagation.

Waves having a longitudinal component (i.e. in the direction of propagation) of magnetic force are called H or TE (transverse electric) waves; those having a longitudinal electric force are E or TM (transverse magnetic) waves.

Various "modes" or configurations of electric and magnetic fields are possible, but the only ones of practical engineering importance are those known as the H_{01} (or TE_{01}) in rectangular guides, and E_{01} (or TM_{01}) and H_{11} (or TE_{11}) in circular guides. Circular guides are generally used only when axial symmetry is required, as in rotating joints.

Rectangular Guide—H_{01} Mode

The lines of electric and magnetic force are shown in Fig. 9.1. The electric field is everywhere perpendicular to the broad face of the guide, and its intensity varies sinusoidally from zero at the sides to a maximum at the middle. The magnetic lines

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of force are closed loops, and the whole pattern moves along the guide, in the direction of propagation, with a velocity known

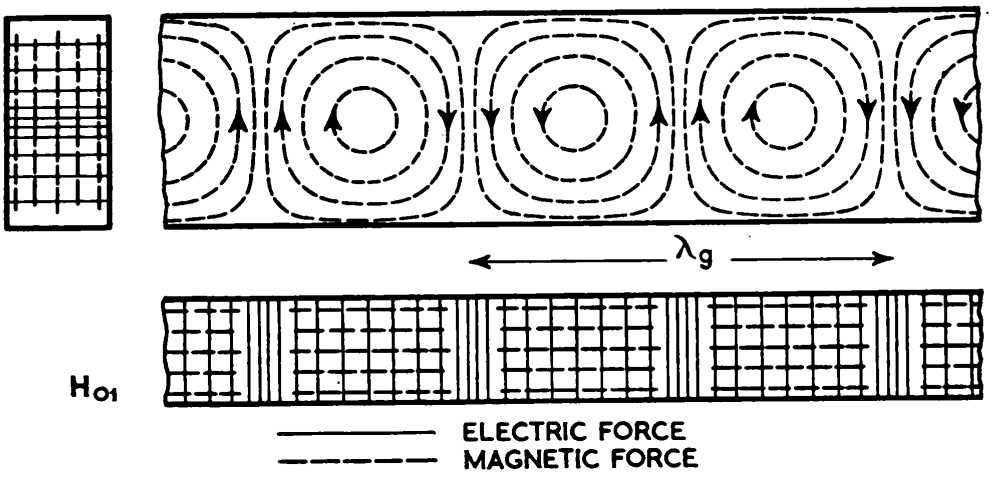

FIG. 9.1.—RECTANGULAR WAVEGUIDE—ELECTRIC AND MAGNETIC FIELDS.

as the phase velocity. At any point, the intensities of both fields vary sinusoidally with time.

Guide Wavelength and Cut-off Frequency

The frequency corresponding to a free-space wavelength equal to twice the wide dimension of the guide is called the cut-off frequency (f_c). Frequencies lower than this cannot be transmitted.

The phase velocity is greater than that of light (c), being $\frac{c}{\sqrt{1-(f/f)^2}}$, where f is the actual frequency of the wave being transmitted.

The wavelength in the guide (λ_g) is therefore greater than in free space (λ_0), the relation between the two being

$$\frac{1}{\lambda_g^2} = \frac{1}{\lambda_0^2} - \frac{1}{\lambda_c^2},$$

where λ_c is the free-space wavelength corresponding to the cut-off frequency of the guide.

Although the field pattern travels along the guide faster than light, it can be shown that intelligence, e.g. a train of waves forming a pulse, is transmitted at a lower velocity, called the group velocity, equal to $c\sqrt{1-(f_c/f)^2}$.

The greater the narrow dimension of the guide, the greater is the amount of power that can be transmitted, but to prevent the formation of other modes, the narrow dimension should be less than half the free-space wavelength of the frequency used.

Many sizes of guide have been used, but common ones are $3 \times 1\frac{1}{2}$ inches for 10 centimetres, and $1 \times \frac{1}{2}$ inch for 3 centimetres. The inside dimensions of these are $2{\cdot}84 \times 1{\cdot}34$ inches

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and 0·9 and 0·4 inch respectively (7·2 × 3·4 and 2·29 × 1·02 centimetres).

Fig. 9.2 shows the lines of current flow in the guide walls. The current is confined to a very thin layer at the inside surface.

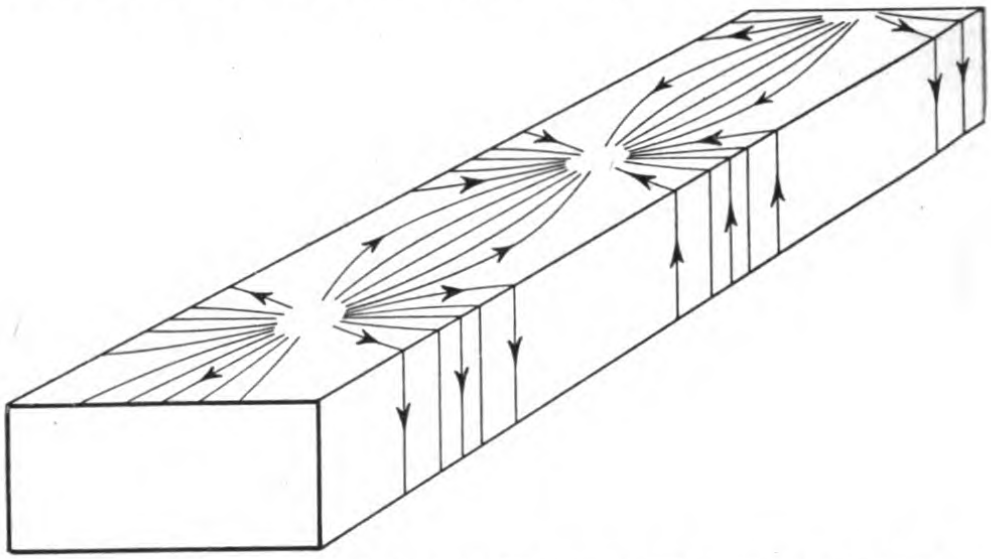

FIG. 9.2.—LINES OF CURRENT FLOW IN WAVEGUIDE WALLS.

Higher Order Modes

If the guide is large enough, modes of transmission can exist wherein the field pattern is repeated more than once across the guide. The resultant patterns are complicated, and in practice the guide dimensions are chosen so that higher modes cannot exist.

Circular Waveguides

The lowest mode in which propagation is possible is the H_{11} mode, the cut-off frequency of which corresponds to a free-space wavelength of $3{\cdot}41a$, where a is the inside radius of the guide.

The next lowest mode is the E_{01} mode, which has a cut-off wavelength of $2{\cdot}61a$. The field patterns for these modes are shown in Fig. 9.3.

The E_{01} mode is used in rotating joints because of its axial symmetry, the diameter of the guide being kept small enough to prevent the transmission of other (unwanted) modes. The H_{11} mode cannot be prevented by this means, and other steps are taken to suppress it.

The formulae relating phase and group velocities, cut-off wavelengths and guide wavelengths are the same as for rectangular guides.

Impedance and Matching

The objections to impedance mismatch, high standing wave ratio, etc., referred to under coaxial lines apply equally to waveguides.

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A difficulty arises in defining what is meant by the impedance of a wavelength because this quantity cannot easily be visualized as a simple ratio of volts/amperes. Various arbitrary definitions have been proposed (generally giving different answers), the principal ones being

$$\text{(i) Wave impedance} = \frac{\text{transverse electric field strength}}{\text{transverse magnetic field strength}}$$

$$= \frac{\eta}{\sqrt{1 - (f_c/f)^2}}$$

The field strengths are measured in volts per metre and amperes per metre, respectively, and the impedance in ohms.

$$\text{(ii) Characteristic impedance} = \frac{(\text{voltage})^2}{\text{power transmitted}}$$

$$= \frac{2b\eta}{a\sqrt{1 - (f_c/f)^2}}.$$

The voltage is the r.m.s. value between opposite mid-points of the broad faces of the guide (assumed rectangular).

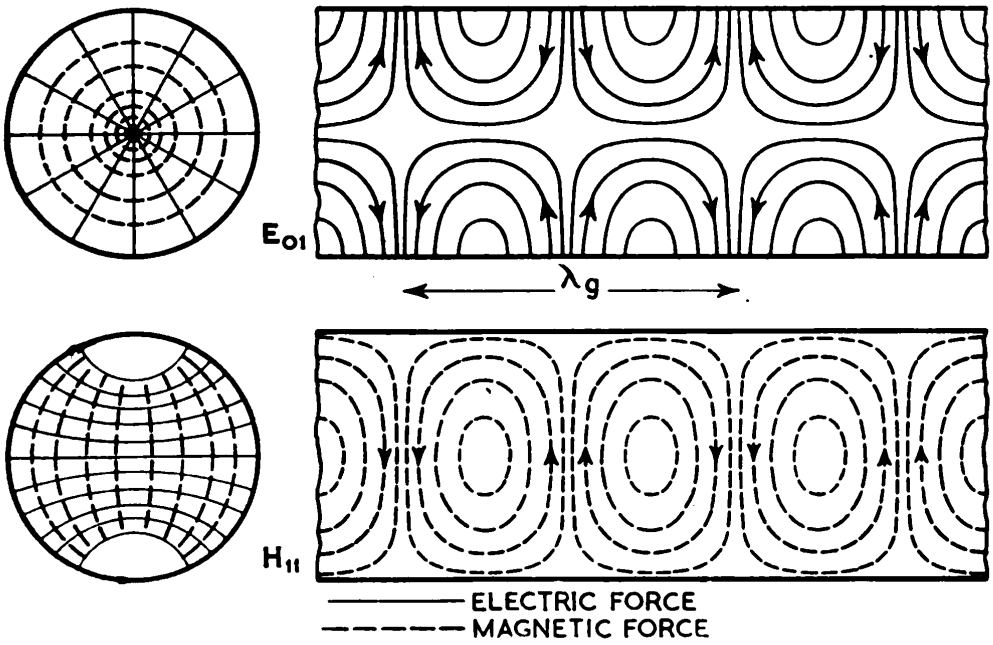

FIG. 9.3.—CIRCULAR WAVEGUIDE—ELECTRIC AND MAGNETIC FIELDS.

In the above expressions, which relate to the H_{01} mode in a rectangular guide, $\eta = 377$ ohms (characteristic impedance of free-space) and a and b are the broad and narrow dimensions (inside) of the guide.

The formulae give, for the two sizes of guide mentioned:

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	Wave Impedance	*Characteristic Impedance*
3 × 1½-inch guide at 10 cm.	524 ohms	495 ohms
1 × ½-inch „ „ 3 „	498 „	443 „

The ambiguity due to different definitions is avoided in practice by using the conception of "normalized" impedance, i.e. impedance is quoted in terms of its ratio to that of the guide. The practical significance of this is that the absence of reflections and therefore standing waves (i.e. S.W.R. = 1) indicates impedance match.

Matching Elements

A metal diaphragm, thin in the direction of propagation, if placed in a waveguide, will act as a lumped impedance and can be used for tuning out unwanted reflections in the same way as inductance or capacitance can be shunted across a two-wire or coaxial transmission line. Fig. 9.4 shows, (*a*) capacitive, (*b*) inductive, (*c*) resonant types, together with their equivalent two-wire circuits.

If the waveguide is carrying appreciable power, the inductive type diaphragm is usually preferred as the capacitive type increases the liability to spark across. Metal posts, inserted in the guide, will also act as impedances.

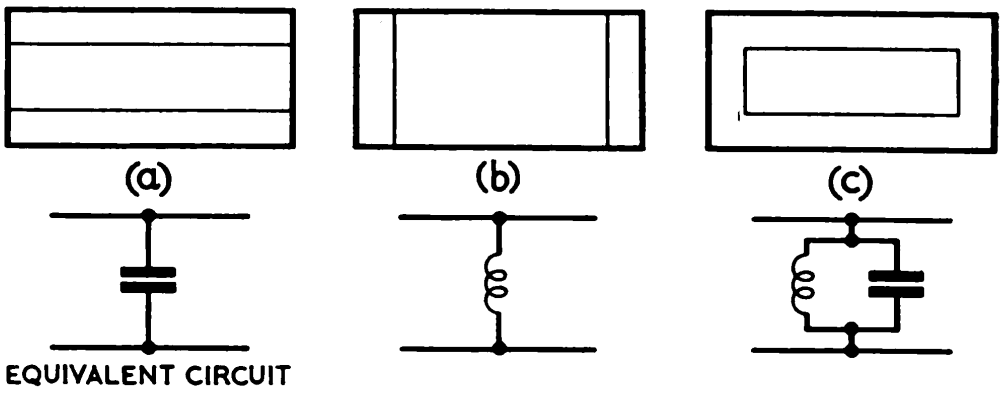

FIG. 9.4.—REACTIVE DIAPHRAGMS FOR WAVEGUIDES.

Where a waveguide is connected to another of different section, impedance matching can be effected by connecting the two by a tapered section, tapering gradually from the cross-section of one guide to that of the other. Alternatively, the transformation may be made by inserting a length of guide of intermediate section, analogous to the quarter-wave transmission line transformer. Similar means can be used to connect a waveguide to a magnetron wherein the output takes the form of a slot in one of the resonant cavities.

A metal plate across a waveguide acts as a short-circuit, but an open circuit is not so easy to obtain with a waveguide because a plain open end will radiate appreciable power.

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As in the case of coaxial lines, stubs fitted with adjustable short-circuiting pistons can be used for matching the load to the waveguide.

In a practical waveguide feeder having discontinuities at several different points, it is necessary to match each discontinuity separately. A single matching element at one end cannot be used because the phase velocity in the guide, and therefore the phase differences between the several points, will vary with change of frequency.

The long line effect mentioned in connection with coaxial feeders is equally applicable to waveguides.

Phase Adjusters and Attenuators

A strip of dielectric material introduced into a waveguide will alter the velocity of the wave and so cause a phase change. The effect is small when the strip is at one side of the guide, and increases to a maximum as it is moved to the centre, where the electric field is most intense.

A piece of lossy material, such as carbon-loaded bakelite, will introduce attenuation if placed in a guide. To avoid reflections, the ends should be tapered. Attenuations up to 20 to 30 decibels can be obtained in this way.

A vane of lossy material can be inserted to an adjustable extent through a slit in the guide to give variable attenuation. If appreciable power is involved, the " vane " may be made of a bent glass tube through which water is circulated.

Power Handling Capacity

The maximum power that can be carried by a waveguide is limited by the dielectric strength of the air inside it, which at normal temperature and pressure is about $2{\cdot}9 \times 10^6$ volts per metre. If $1{\cdot}5 \times 10^6$ (peak) be taken as a safe working figure, the relation between power and voltage contained in the formula for characteristic impedance shows that a $3 \times 1\frac{1}{2}$-inch standard guide can carry 2·6 megawatts at 10 centimetres and a $1 \times \frac{1}{2}$-inch guide 265 kilowatts at 3 centimetres.

If the S.W.R. is considerably greater than unity the maximum permissible power will be less ; for example, a S.W.R. of 1·5 will reduce the power by approximately 30 per cent, assuming the peak voltage is kept to a specified figure.

In aircraft at great altitudes the maximum permissible power will be approximately proportional to the pressure of the air. When it is desired to transmit the greatest possible power, waveguides are sometimes pressurized.

Roughness and sharp points will reduce the maximum permissible voltage.

Attenuation in Waveguides

The currents in the walls of the guide cause losses due to ohmic resistance of the material.

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For a rectangular guide transmitting the H_{01} mode the attenuation is

$$\alpha = \frac{R_s}{b\eta_0}\left[\frac{1+\frac{2bf_c^2}{af^2}}{\sqrt{1-(f_c/f)^2}}\right] \text{ nepers per metre (multiply by 8·7 to get decibels)}$$

where R_s is the surface resistivity of the guide material,
η_0 the intrinsic impedance of free space,
f the working frequency,
f_c the cut-off frequency of the guide,
a and b the broad and narrow (internal) dimensions.

$R_s = \sqrt{\frac{w\mu_m}{2\sigma_m}}$, where μ_m is the permeability and σ_m the conductivity of the guide material.

For copper, taking $\mu_m = 4\pi \times 10^{-7}$, $\sigma_m = 5{\cdot}8 \times 10^7$, this formula gives an attenuation of approximately 2 decibels per

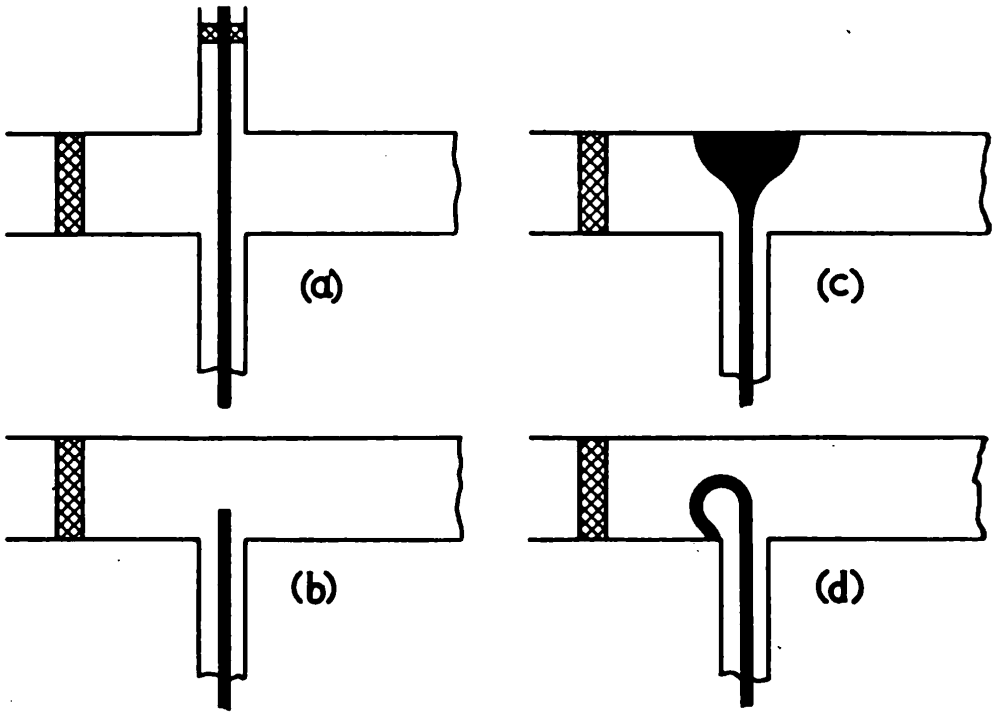

FIG. 9.5.—METHODS OF COUPLING COAXIAL FEEDER TO WAVEGUIDE.

100 metres of 3 × 1½-inch guide at 10 centimetres, and 11 decibels per 100 metres for 1 × ½-inch guide at 3 centimetres. The attenuation will be greater if the interior surface of the guide is dirty or oxidized.

Connection between Concentric Feeder and Waveguide

Various methods are shown in Fig. 9.5. (*a*) shows the inner conductor of the concentric line carried right across the wave-

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guide and terminated in a quarter-wave stub. Matching is effected by adjusting the length of this stub and also the plunger closing the left-hand end of the guide. In (*b*) the inner conductor terminates in a probe, matching being by varying the amount of penetration into the guide, and adjustment of the short-circuiting plunger as before. (*c*) is known as the " door-knob " type, and (*d*) is a coupling loop.

Waveguide Couplings

The choke coupling shown in Fig. 9.6 is used when two lengths of waveguide are to be coupled together. *B* is a quarter-wave slot, presenting infinite impedance at the point *C* and therefore giving zero impedance across the gap *A*, which is one quarter

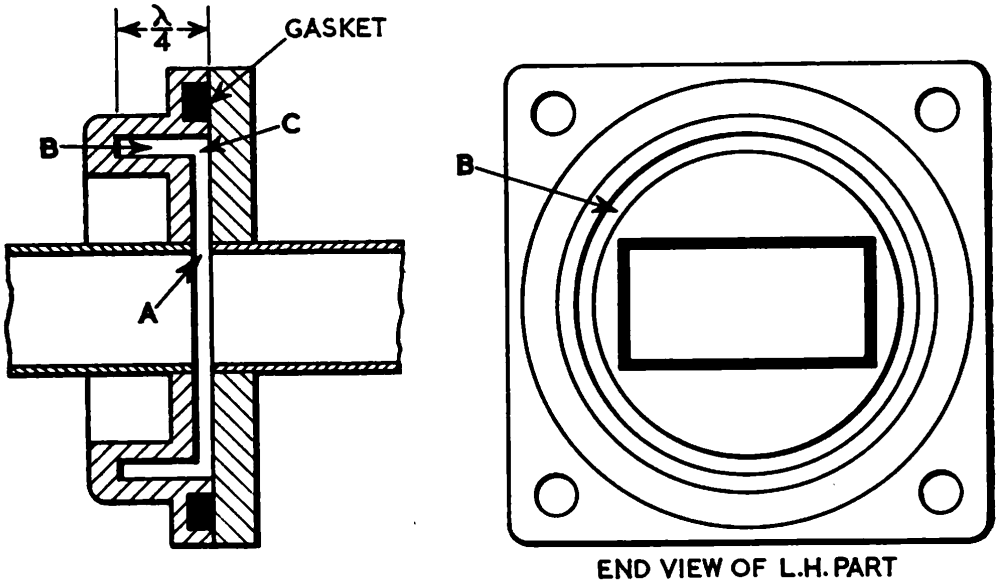

FIG. 9.6.—CHOKE JOINT FOR WAVEGUIDE.

wavelength from *C*. The two lengths of guide are thus effectively connected without relying on actual contact at this point.

The outline of the slot should theoretically follow that of the guide cross-section, but in practice it is made circular (for ease of manufacture), of radius calculated to give the best overall result, having regard to the fact that the current flow is a maximum at the centre of the broad face of the guide.

Rotating Joint

A common type is shown in Fig. 9.7. The H wave in the main rectangular guide excites an E wave in the circular part. As the E wave has axial symmetry, the top part of the assembly can be rotated relatively to the lower part without adverse effects.

Coupling between the two halves of the circular part is by making use of the quarter-wave principle as in the case of

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concentric lines. Excitation of the unwanted H_{11} wave in the circular part is prevented by resonant filter rings.

An alternative method of making a rotating joint is to convert from waveguide to concentric feeder and back again, the rotating part being in the concentric part.

Waveguide Bends

If space permits, a long gradual bend (several wavelengths long) is best. If it is necessary to economize space, bends of

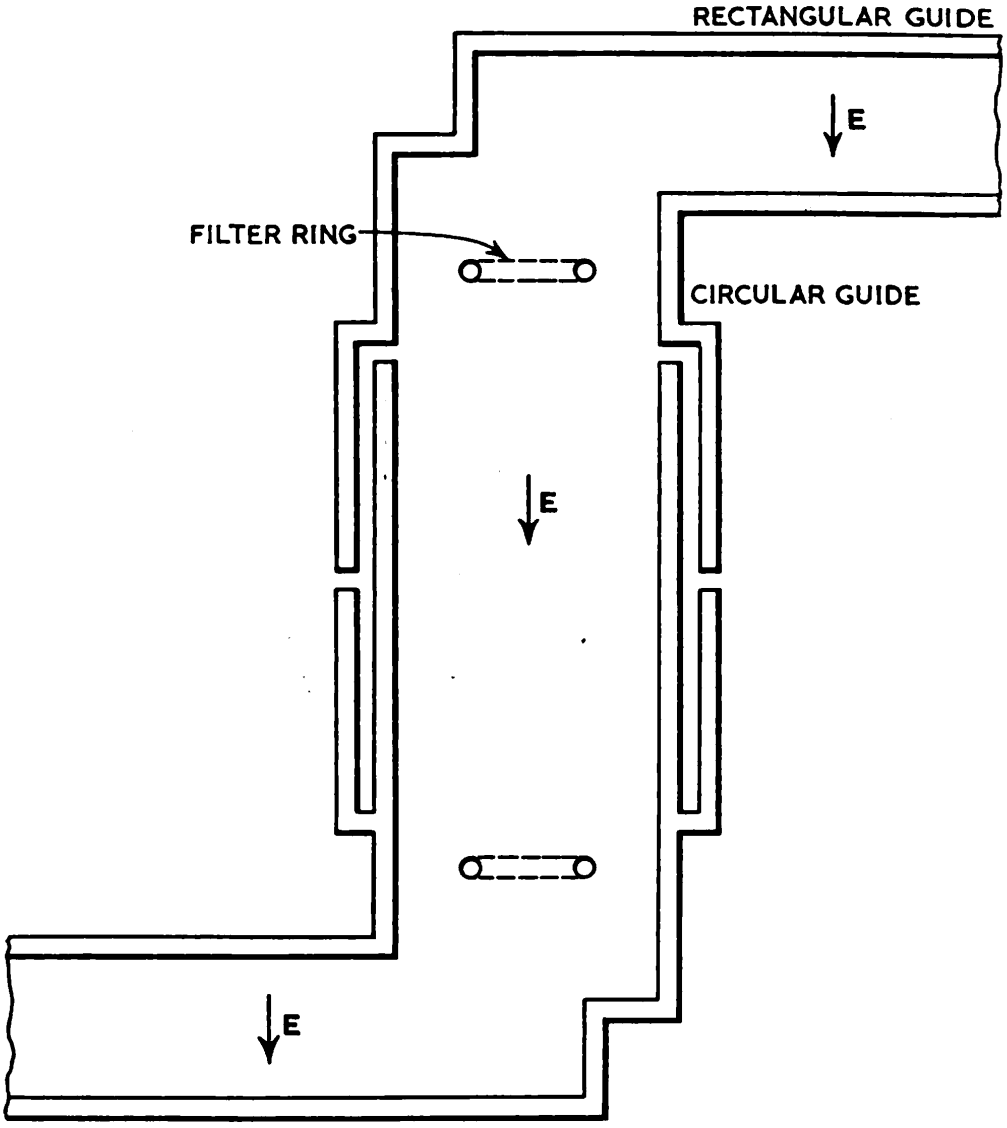

FIG. 9.7.—ROTATING JOINT FOR WAVEGUIDE.

the types shown in Fig. 9.8 can be used. If properly designed, these can be used over a frequency band of about ± 10 per cent of mid-frequency.

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Waveguide Junctions

If power is fed into arm A of a tee-junction (Fig. 9.9) made on the broad face of a guide, the power will be equally divided

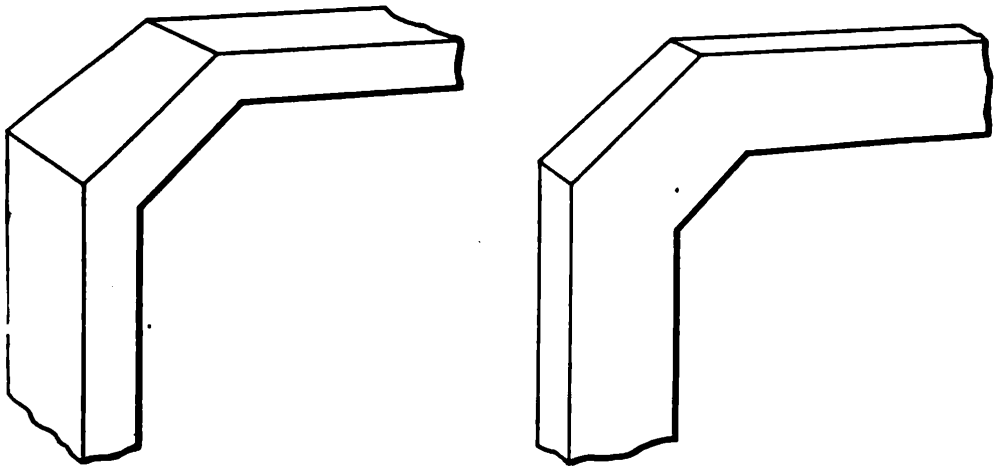

FIG. 9.8.—WAVEGUIDE BENDS.

between arms B and C. The directions of the lines of electric force in the neighbourhood of the junction indicate that the waves in B and C will be in opposite phase. This type is known as a series tee or series junction.

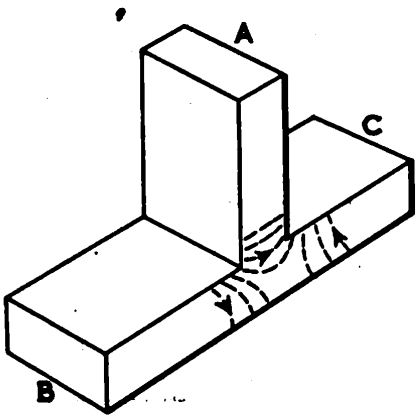

FIG. 9.9.—SERIES TEE JUNCTION FOR WAVEGUIDE.

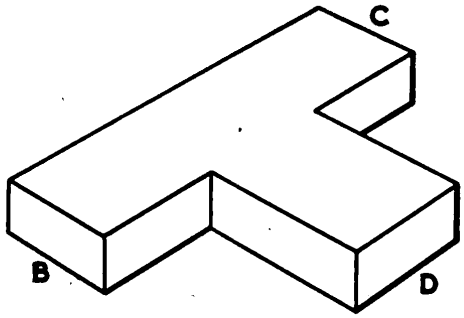

FIG. 9.10.—SHUNT TEE JUNCTION FOR WAVEGUIDE.

Fig. 9.10 shows a shunt junction; in this case the waves in branches B and C are in phase when power is fed into D.

Magic Tee

This, shown in Fig. 9.11, is a combination of the two types of junction, and has certain useful properties.

Power fed into A will give waves in B and C in opposite phase, whereas if the power is fed into D the outputs in B and C will be in phase. Also, power fed into A will not produce any power

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flow in D, or vice versa. Arms A and D are said to be conjugate, and this type of coupling is sometimes called a bridge.

Rat Race

This is shown in Fig. 9.12 and consists of a circular ring of rectangular waveguide with branches spaced as shown. The

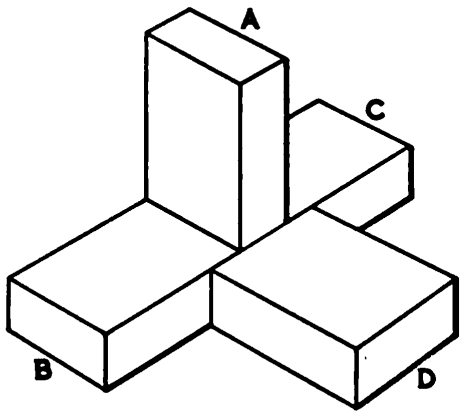

FIG. 9.11.—MAGIC TEE.

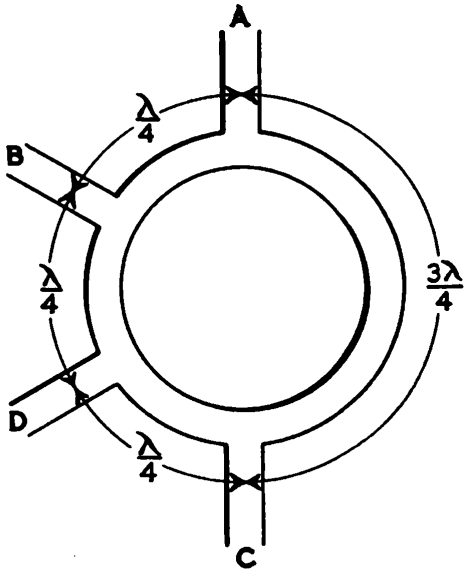

FIG. 9.12.—RAT RACE.

branches are lettered to correspond with Fig. 9.11 and it will be seen that the device is equivalent to the magic tee.

A useful application of the rat race is as a duplexer for a common T–R circuit (see Chapter X).

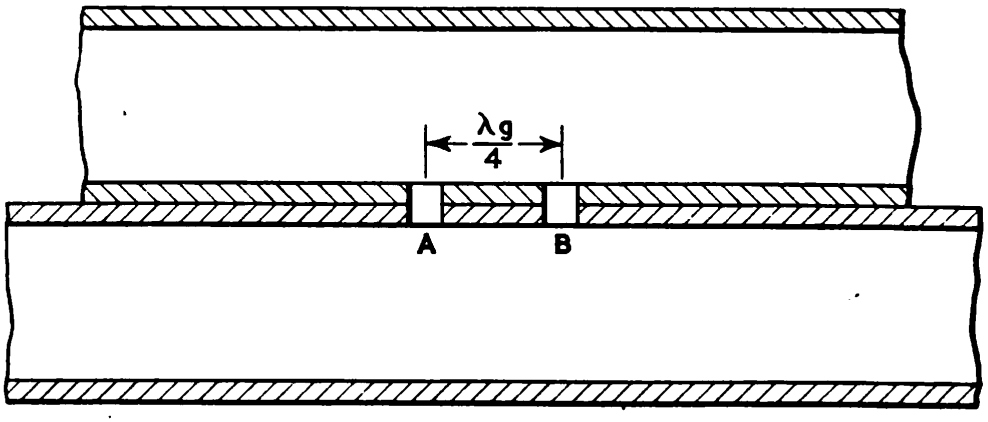

FIG. 9.13.—DIRECTIONAL COUPLER—TWO-HOLE TYPE.

Directional Couplers

These are used for transferring a fraction of the power in one waveguide to another guide in such a way that the wave excited in the second guide travels only in one direction.

Coupling is defined as the ratio of the amplitudes of the waves

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in the main and auxiliary guides, and directivity is the ratio of the wave in the wanted direction in the auxiliary guide to that in the unwanted direction.

In the two-hole type, shown in Fig. 9.13, the waves excited

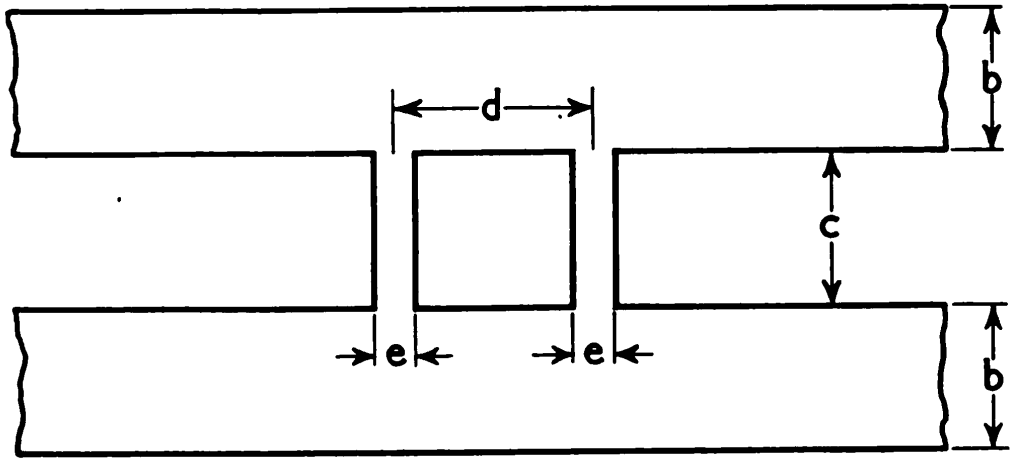

FIG. 9.14.—DIRECTIONAL COUPLER—DOUBLE STUB TYPE.

in the auxiliary guide, through the two holes, reinforce in the wanted direction and cancel in the opposite. The arrangement is necessarily somewhat frequency sensitive.

The double stub coupler, shown in Fig. 9.14, gives tighter

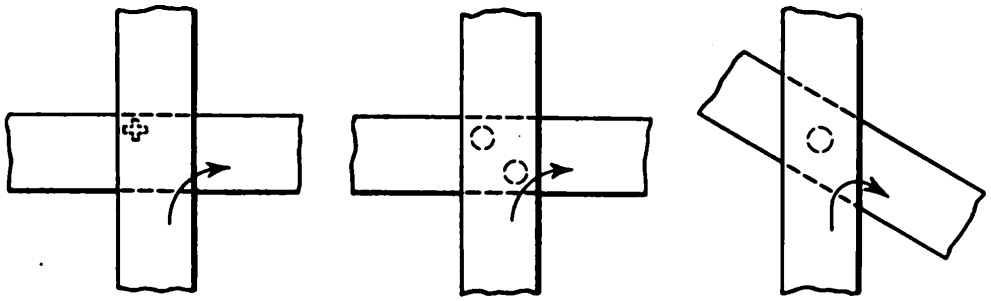

FIG. 9.15.—DIRECTIONAL COUPLERS—VARIOUS TYPES.

coupling and less variation with frequency than the simple two-hole type. The dimensions *c*, *d* and *e* depend on the coupling required; approximate values (centimetres) for two values of coupling for 3 × 1½-inch guide working at 10 centimetres are:

Coupling	*c*	*d*	*e*
20 db.	2·9	3·48	0·36
10 db.	2·5	3·55	1·2

A change of frequency of 5 per cent will only alter the coupling by about 1 decibel. The directivity is about 10–20 decibels.

In general, excitation through a coupling hole is due partly to magnetic and partly to electric coupling. By choosing suitable dimensions and positions for the hole or holes the two

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effects may be made to cancel in the unwanted direction and reinforce in the other. Many variations are possible, some examples being shown in Fig. 9.15.

Where one end of the auxiliary guide is " blind " a wedge of lossy material should be inserted to absorb any power in the unwanted direction to prevent it being reflected.

Horns

A waveguide required to act as a radiator can be approximately matched to free space by being flared at the open end. The widening may be in the E-plane, or the H-plane, or in both. The first two are called sectoral horns, the third pyramidal.

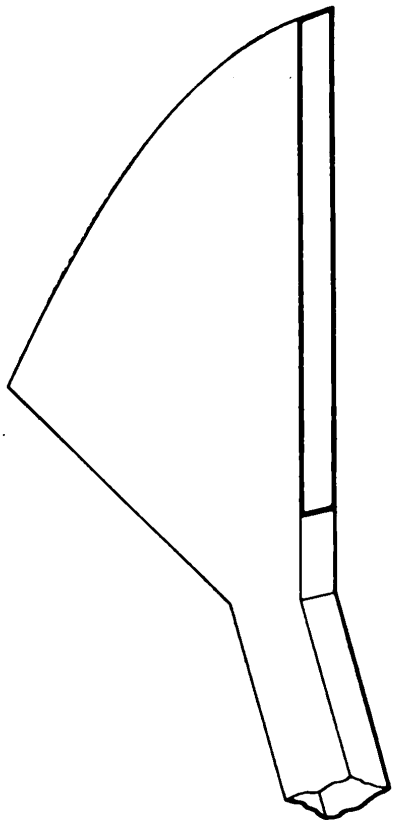

FIG. 9.16.—HOGHORN.

The angle of flare should not be too great, and this limits the area of aperture obtainable with a horn of reasonable length. For this reason a horn is usually used in association with a parabolic mirror in order to obtain a reasonably narrow beam.

The hoghorn, shown in Fig. 9.16, gives a beam which can be made from about 5 to 15 degrees wide in the vertical plane and 20 to 30 in the horizontal. It can be used for feeding a cheese aerial, the combination giving a fan-shaped beam narrow in the horizontal plane and relatively wide in the vertical. This type is commonly used on marine radars where accurate horizontal resolution is required, but some latitude in the vertical is desired to offset effects of the ship pitching and rolling.

Solid Wire Guides

If the end of a coaxial cable is opened out into the form of a horn, and the inner conductor continued as shown in Fig. 9.17,

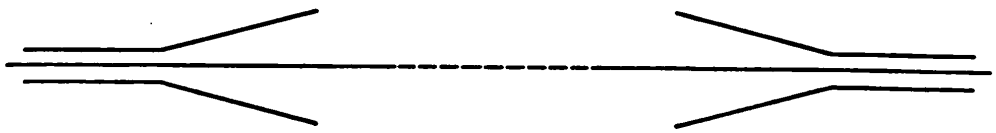

FIG. 9.17.—SOLID WIRE GUIDE.
Showing method of launching wave.

this conductor is able to support a surface wave and thus act as a waveguide.

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The fields round the guide extend theoretically to infinity, but if the phase velocity along the wire is somehow reduced below its free space value it is found that most of the energy is conveyed inside an imaginary cylinder of reasonable diameter, concentric with the wire.

The necessary reduction of phase velocity, a few per cent is sufficient, can be obtained by modifying the surface of the wire in some way; an easy method is to coat the wire with a thin layer of enamel. The attenuation is comparable with that of conventional waveguides. So far, this type of guide only appears to have been used experimentally.

Cavity Resonators

If a length of waveguide is closed at each end, and oscillations are excited at the appropriate frequency, the waves reflected

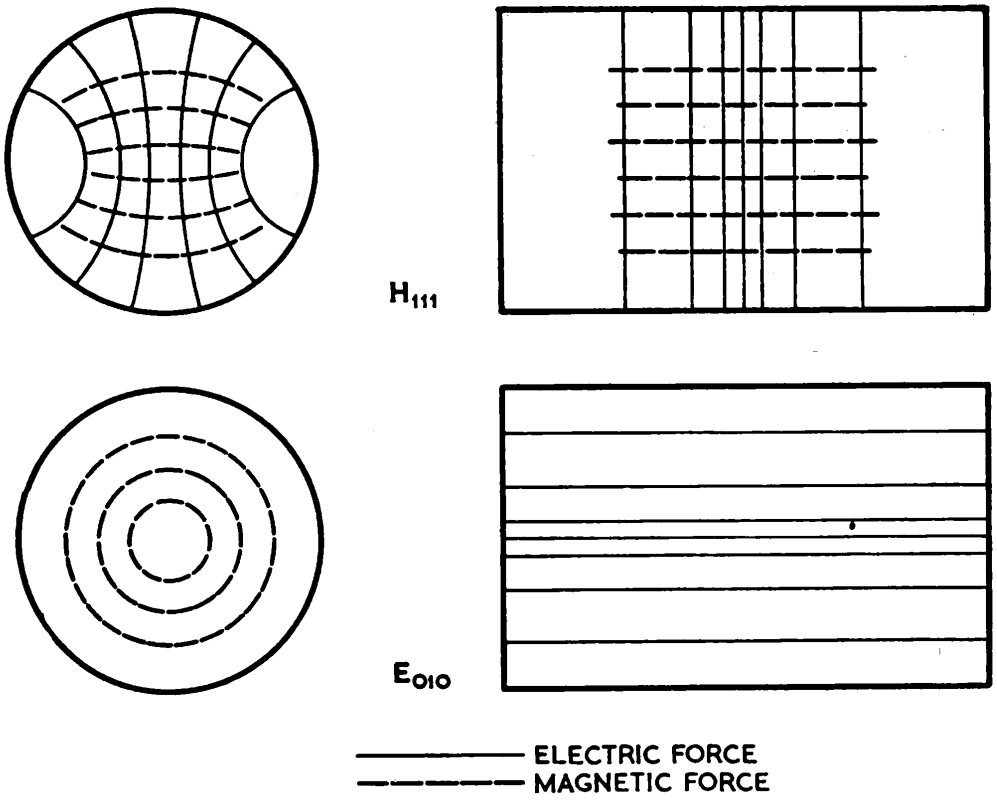

FIG. 9.18.—CAVITY RESONATOR.
Cylindrical type.

from each end will set up standing waves, and the cavity will act as a resonator.

Oscillation may take place in modes corresponding to the H or E waves in waveguides, and either of these two main types can occur in a number of modes, each of which corresponds to a different frequency.

In practice it is frequently found that a cavity, or echo box

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as it is often called, can oscillate in two or three modes whose frequencies are very close together, and it is sometimes a matter of some difficulty to design an echo box for a particular wavelength in such a way that the likelihood of this happening is small.

For ease of manufacture, echo boxes are usually made of cylindrical shape, with one end movable for adjustment of wavelength.

The lowest modes of oscillation of cylindrical cavities are the H_{111} and E_{010} modes, the field patterns of which are shown in Fig. 9.18. The wavelengths are

$$H_{111} \text{ mode}: \quad \lambda = \frac{1}{\sqrt{\frac{0{\cdot}0859}{a^2} + \frac{0{\cdot}25}{l^2}}}$$

$$E_{010} \text{ mode}: \quad \lambda = 2{\cdot}61a$$

where a and l are the radius and length of the cylinder, respectively.

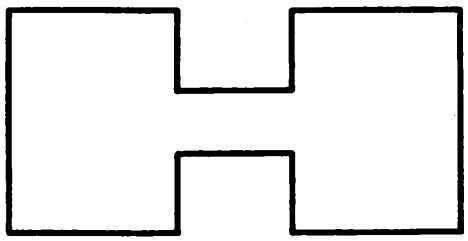

FIG. 9.19.—CAVITY RESONATOR. Re-entrant type.

For special purposes, e.g. the klystron or the T.R. cell, re-entrant cavities are used. One form is shown in Fig. 9.19.

The Q of a cavity is very high compared with that of a conventional tuned circuit consisting of lumped inductance and capacitance, values of 10,000 to 50,000 being obtainable in practice.

Resonant cavities may be excited by a coupling loop or a probe, or, as in the case of the klystron, by an electron stream. Power can be taken out in a similar manner.

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CHAPTER X

AERIAL SYSTEMS AND RADIATION

THE basic aerial is usually a dipole. By itself, this does not possess any marked directional properties, and it is necessary to use an array of several dipoles or to provide a mirror for concentrating the radiated energy in the required direction. The former method is applicable to metre wavelengths, and the latter to centimetric.

In present-day radars it is common practice to use the same aerial for transmission and reception.

Polar Diagrams

The polar diagram indicates the manner in which the field strength, at a given distance, varies with direction. In the example shown in Fig. 10.1 the relative field strengths at the points *A* and *B* (supposed to be equidistant from the aerial

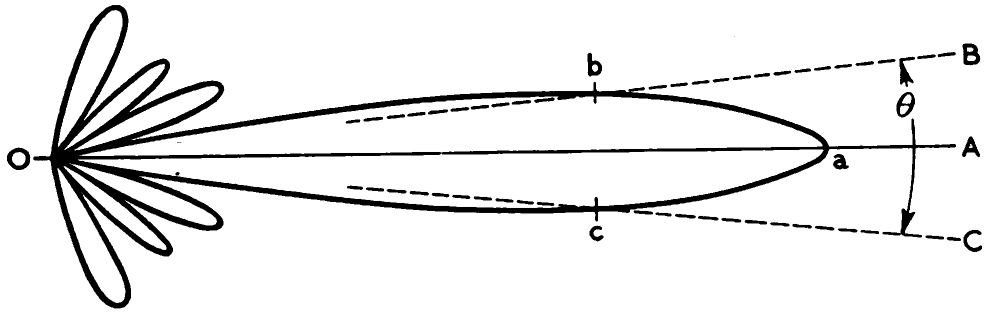

FIG. 10.1.—EXAMPLE OF AERIAL POLAR DIAGRAM.

at *O*) are proportional to the lengths *Oa* and *Ob*, respectively. Diagrams can be drawn for both the horizontal and vertical planes.

It follows from the principle of reciprocity that the polar diagram of an aerial is the same whether it is used for transmission or reception.

Side Lobes

In practical aerial systems, designed to produce narrow beams, it is impossible to avoid the production of auxiliary beams called side lobes. These are shown in Fig. 10.1 ; their number and magnitudes depend on the particular aerial considered.

To avoid confusion due to echoes from side lobes, these must be made as small as possible by careful design of the aerial.

Special Polar Diagrams

By using a suitably shaped mirror, an aerial system may be made with a special polar diagram for some particular purpose.

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An example is shown in Fig. 10.2, where the field strength at a point P is represented by the length of the line Op, which is proportional to the cosecant of the angle of elevation (θ) within the limits of coverage (i.e. between OA and OB). Within these

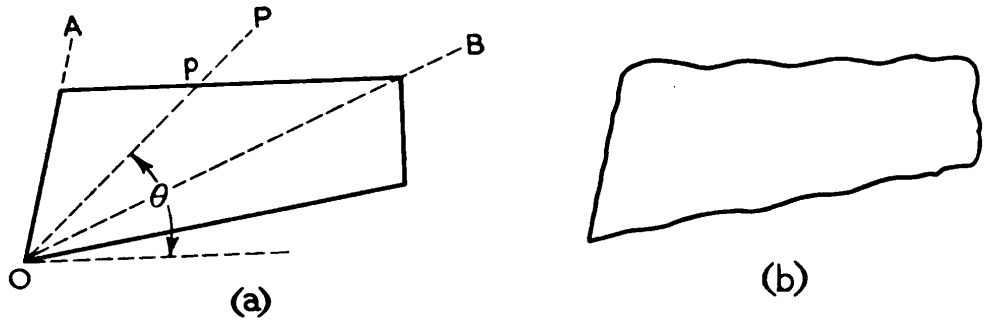

FIG. 10.2.—COSEC$^2\theta$ POLAR DIAGRAM.
(a) Theoretical. (b) Typical practical.

limits this aerial will give echoes of constant strength from an aircraft approaching at constant height.

An aerial of this type is called a cosec$^2\theta$ aerial, and is suitable for early warning radars. The practical diagram is more like Fig. 10.2 (b).

The same type of diagram, inverted, is suitable for an airborne P.P.I. display.

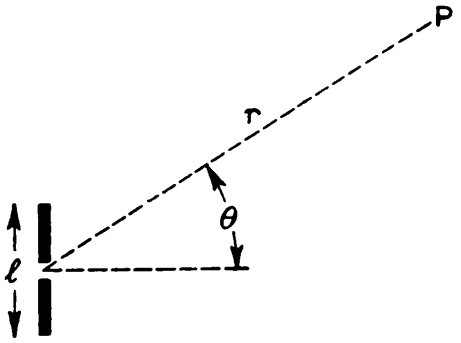

FIG. 10.3.—DIPOLE.

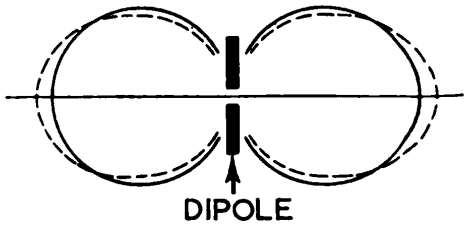

FIG. 10.4.—POLAR DIAGRAM OF DIPOLE.
Full line—Short dipole.
Dotted line—Half wave dipole.

Radiation from Short Dipole

If a short dipole, of length l (small compared with the wavelength), is centre fed with a current of I amperes (see Fig. 10.3), the electric and magnetic fields at a point P, distant r, are

$$E = \frac{\omega Il}{8\pi\epsilon c^2 r}\cos\theta = \frac{\omega Il}{2\times 10^7 r}\cos\theta \quad \begin{matrix}[\text{taking velocity } (c)\\ \text{as } 3\times 10^8]\end{matrix}$$

$$H = \frac{\omega Il}{8\pi rc}\cos\theta = \frac{\omega Il}{24\times 10^8\pi r}\cos\theta.$$

(I, E and H are r.m.s. values.)

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The dipole is not an isotropic radiator, i.e. it does not radiate equally in all directions. The average energy crossing unit area at point P is

$$EH = \frac{\omega^2 I^2 l^2}{48 \times 10^{15} \pi r^2} \cos^2 \theta$$

and the total energy radiated is

$$\frac{\omega^2 I^2 l^2}{18 \times 10^{15}} = \frac{20\pi^2 I^2 l^2}{\lambda^2} = R_o I^2,$$

where $R_o = 20\pi^2(l^2/\lambda^2)$, which is called the radiation resistance of the aerial.

For a half-wave dipole ($l = \lambda/2$) the fields can be obtained by replacing $\cos\theta$ in the expressions for E and H by $\dfrac{4\cos\left(\frac{\pi}{2}\sin\theta\right)}{\pi\cos\theta}$.

For an infinitely thin half-wave dipole the radiation resistance is 73 ohms. If the diameter is 1/100 of the length, the figure is about 20 per cent less. Due to end effects, the actual physical length of a half-wave dipole is about 5 to 10 per cent less than the theoretical.

Bandwidth of Dipole

This is the total frequency range over which the power radiated for a given input voltage is equal to or greater than half the maximum, i.e. the power at resonance. A thick dipole has a lower effective inductance and therefore a greater bandwidth than a thin one. Accurate calculation is complex, but approximate figures for the bandwidth of dipoles of length/diameter of 100 and 10,000 are 15 and 8 per cent respectively.

Polar Diagram of Dipole

Fig. 10.4 shows the form of the vertical polar diagram for a vertical dipole in free space. The full lines (which are circles) apply to a short dipole and the dotted lines to a half-wave dipole.

From considerations of symmetry it is obvious that the horizontal polar diagram of a vertical dipole is a circle having its centre at the dipole.

Dipole Arrays (for Metre Wavelengths)

To obtain a narrow beam an array as shown in Fig. 10.5 is used, the number of dipoles and their spacing depending on the sharpness required. This arrangement gives a

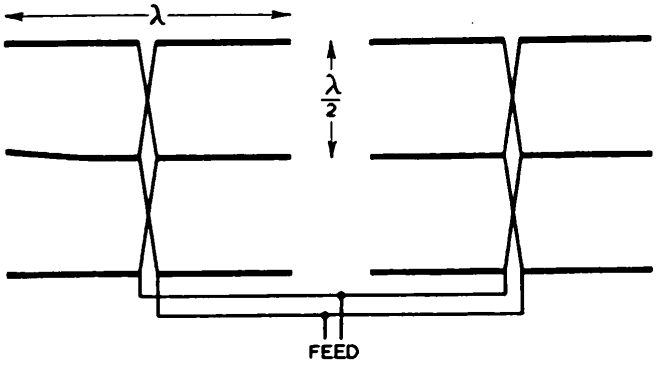

FIG. 10.5.—DIPOLE ARRAY.
Showing method of feeding.

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polar diagram something like that shown in Fig. 10.1, but with similar backward radiation unless a sheet of metal or wire netting is placed behind the array to eliminate this.

The Folded Dipole

If the radiation resistance of a half-wave dipole (60 to 70 ohms) is too low to be matched to a convenient feeder, a folded dipole (Fig. 10.6) can be used.

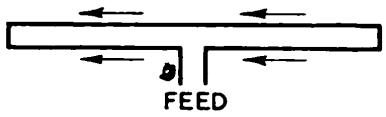

FIG. 10.6.—FOLDED DIPOLE.

The feed is taken to one dipole, and the second may be regarded as voltage-fed across its ends from the first, whence the currents in the two are in the same direction as shown. So far as radiation is concerned, the current is effectively doubled and the power quadrupled, whence the radiation resistance is quadrupled since it is referred to the input current.

The Yagi Aerial (Fig. 10.7)

This consists of a centre-fed dipole $\lambda/2$ long, with from about 3 to 10 auxiliary dipoles (called directors) in front of it, and one (reflector) behind. The directors are shorter than the main dipole, whilst the reflector is longer.

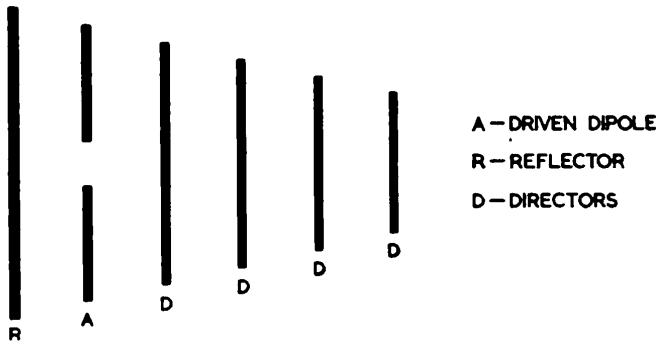

FIG. 10.7.—YAGI ARRAY.

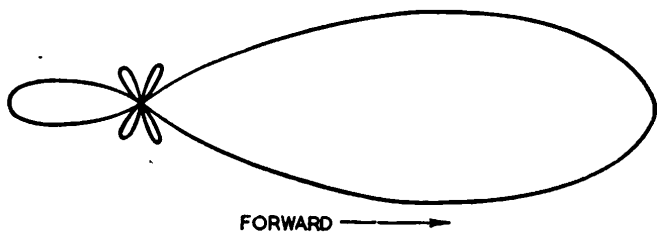

FIG. 10.8.—POLAR DIAGRAM OF YAGI ARRAY.

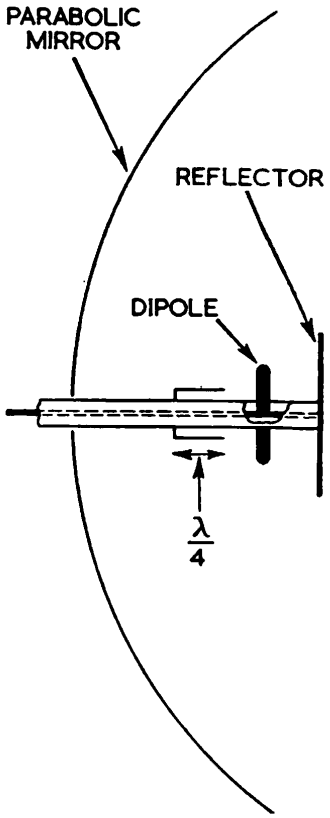

FIG. 10.9.—DIPOLE IN PARABOLIC MIRROR.

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Currents induced in directors and reflectors modify the field produced by the main dipole and a polar diagram of the type shown in Fig. 10.8 is obtained.

The Yagi array has been extensively used for wavelengths of about 1 to 2 metres. It can conveniently be made as an all-metal structure, with a folded dipole as the active member.

Dipole in Parabolic Mirror (Fig. 10.9)

This is a very common aerial arrangement for centimetric wavelengths. The dipole is placed at the focus of the mirror and a sharp beam is obtained if the mirror diameter is several wavelengths. The polar diagram is of the type shown in Fig. 10.1. In this diagram b and c are the " half-power " points ($Ob^2 = Oc^2 = \frac{1}{2}Oa^2$). The beam width ($\theta$) to half-power points is approximately λ/D radians, where D is the diameter of the mirror and λ the wavelength. The side lobes (indicated arbitrarily in the diagram) vary in strength and position according to the details of the particular case.

As an example, the theoretical beam width obtained with a mirror 2 metres diameter, at 10-centimetre wavelength, is 2·9°, but the actual beam will be wider, due to practical imperfections.

The dipole is fed by a concentric line, with unbalance to balance converter as shown. A small parasitic dipole or a circular reflector prevents direct forward radiation from the dipole.

Waveguide Feed

The dipole of Fig. 10.9 may be replaced by the open end of a waveguide, as in Figs. 10.10 (a) and (b). The " open " end

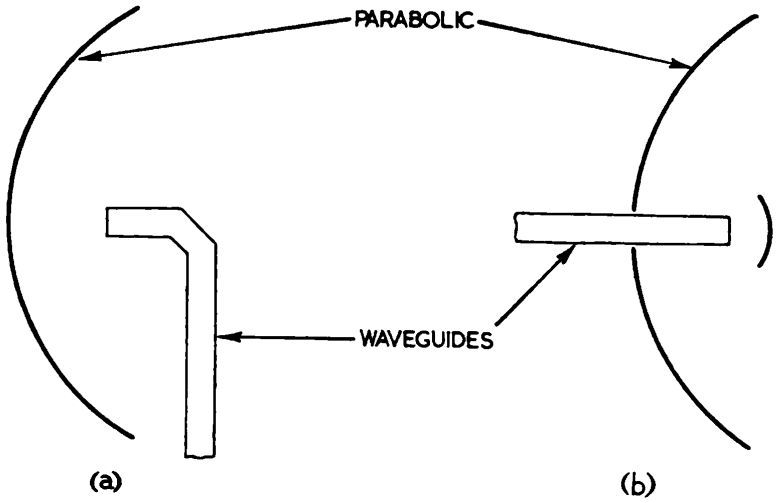

FIG. 10.10. — PARABOLIC MIRROR FED BY WAVEGUIDE.
(a) Front feed.
(b) Back feed.

of the guide is closed by an insulating window to exclude dirt and moisture.

Cheese Aerials

Fig. 10.11 shows a " cheese " aerial fed by a waveguide with a horn termination. It consists of a section of parabolic mirror

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closed in at top and bottom with flat metal plates. This aerial gives a beam which is narrow in the horizontal plane and relatively wide in the vertical. It is suitable for use on ships to avoid losing the target when the ship rolls or pitches.

The horizontal and vertical beam widths (in degrees) are roughly $100(\lambda/b)$ and $100(\lambda/h)$ respectively, where b is the breadth and h the height of the aperture. Side lobes may be about 5 per cent of the main beam in strength.

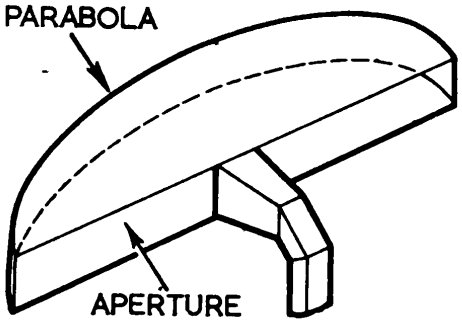

FIG. 10.11.—CHEESE AERIAL.

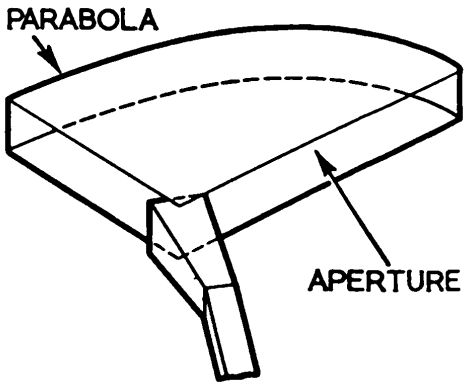

FIG. 10.12.—HALF-CHEESE.

Half-cheese Aerial

This is shown in Fig. 10.12 and is a normal cheese with one half cut off and the feed slewed round to direct the energy into the half which remains.

This aerial gives a better polar diagram than the normal full cheese, the side lobes being reduced, largely because the shadow effect of the feed is practically eliminated instead of coming in the centre of the beam.

Slot Aerials

A slot in an infinite metal plate (in practice this means a large plate), if excited by an oscillator between mid-points of its sides (Fig. 10.13), will radiate with the same polar diagram as a normal dipole of the same shape and size as the slot, except that the planes of electric and magnetic force will be interchanged. A vertical slot therefore gives horizontal polarization.

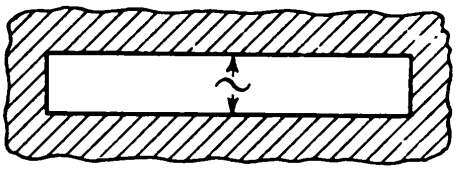

FIG. 10.13.—SLOT AERIAL.

Slot Arrays

Slots cut in either the narrow or wide face of a rectangular waveguide (Fig. 10.14) and spaced half a guide wavelength apart, will interrupt the lines of current flow in the walls of the guide and act as radiators. The necessary phase reversal between successive slots (to compensate for the fact that they are half

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a wavelength apart) is obtained either by staggering or inclining, according to whether the wide or narrow face of the guide is used. The former arrangement gives polarization perpendicular to the length of the guide and the latter parallel to this. The beam is narrow in the plane containing the guide and wide perpendicular to it.

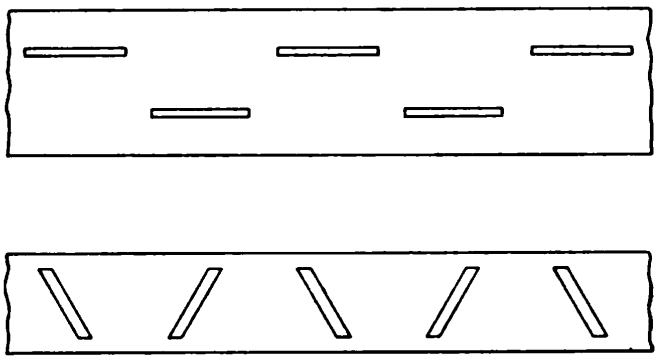

FIG. 10.14.—SLOT ARRAYS.

If the phase velocity in the guide is varied by moving a strip of dielectric in the guide, the effect is to swing the beam.

Polar Diagram of Parabola

For a parabolic reflector which is uniformly "illuminated" over the whole of its area the width of the main beam is approximately $60(\lambda/D)$ degrees to half-power points and $140(\lambda/D)$ to first zeros. These relations hold good for diameters (D) greater than about 5 wavelengths.

A practical dipole or other radiator will not, however, give uniform illumination over the whole area of the mirror, that at the edges being less than in the centre. The result of this is that, in practice, the beam will be some 20 to 30 per cent wider than is indicated by the above formulae. There is also some reduction in side lobes, which is advantageous.

Power Gain and Equivalent Area of Aerial

The power gain of a transmitting aerial is the ratio of the power flow across an element of area, in the direction of maximum radiation, to what it would be if the aerial radiated the same total power equally in all directions.

The power gain of a half-wave dipole is 1·63 and that of a Hertzian dipole (a dipole short compared with λ) is 1·5.

The theoretical power gain of a uniformly illuminated circular parabolic mirror is $(\pi D/\lambda)^2$.

The equivalent or effective receiving area of an aerial is the ratio of the power collected by an aerial (assumed to be pointed in the direction of maximum pick-up) and delivered into a matched load to the power flowing across unit area at the position of the aerial, assuming the presence of the aerial does not disturb the wave.

By the principle of reciprocity, the gain of an aerial is the same whether it is used for transmission or reception. It can

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be shown that the gain of an aerial is equal to $\frac{4\pi A}{\lambda^2}$, where A is the equivalent receiving area. For the ideal case of the uniformly illuminated parabola this formula becomes $(\pi D/\lambda)^2$ as before.

For practical systems the equivalent area is only about 60 per cent of the theoretical (i.e. physical) area. On this basis, the term KA_tA_r in the basic radar equation (Chapter II) becomes $0{\cdot}09\pi^3D^4 = 2{\cdot}8D^4$ for the case of a common aerial system having parabolic mirrors of diameter D metres, and the basic radar equation for free space reduces to

$$\text{range (kilometres)} = 0{\cdot}0005\frac{D}{\sqrt{\lambda}}\sqrt[4]{\frac{W_t}{W_r}}$$

(assuming an effective echoing area of 3·5 square metres for the target).

Determination of Bearing

From Fig. 10.15 it can be seen that the variation of signal strength with bearing is small in the neighbourhood of the maximum, and although a bearing determined by swinging the aerial until maximum strength is obtained may be accurate enough for search radars (especially centimetric with reasonably

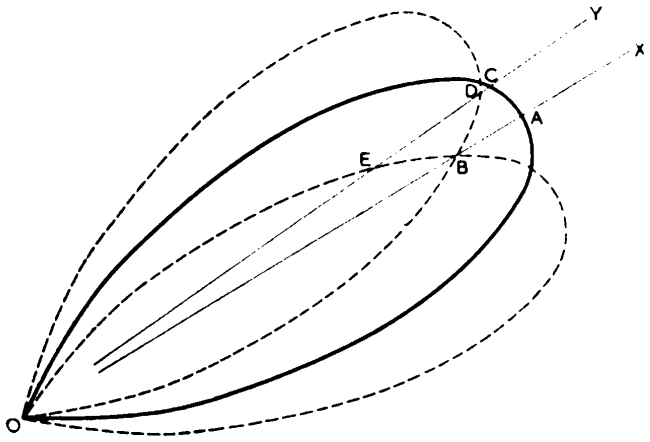

FIG. 10.15.—PRINCIPLE OF "SPLIT."

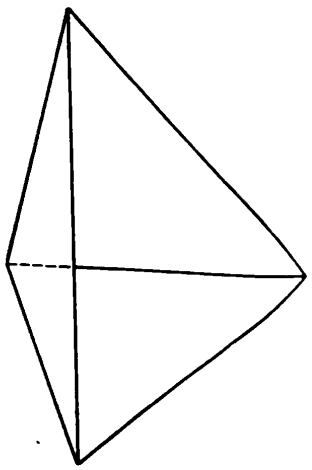

FIG. 10.16.—CORNER REFLECTOR.

narrow beams), this method is not good enough when really accurate results are required.

If the aerial is swung through a small angle on each side of a mid-position it will be seen that, whereas a target at X will give equal responses for the two positions of the beam, one at Y will give signals whose strengths differ considerably. This method, called "split", affords a means of determining the position of maximum signal, and therefore the bearing, accurately.

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Determination of Elevation

With centimetric radars where the beam can be made narrow enough to avoid (except at very small elevations) the effect of ground reflections, elevation can be determined in the same way as bearing.

With older radars, using metre waves, the difficulty of obtaining really narrow beams made it impossible to avoid the effect of ground reflections.

If the earth were a perfect conductor, signals received on two horizontal dipoles at different heights, one above the other, would be in phase but would differ in magnitude. Comparison of the two signal strengths by means of a goniometer or other device would afford a means of calculating the elevation. As the earth is not a perfect conductor, this method would suffer from inaccuracies. The expedient of using an artificial earth consisting of a mat of wire-netting surrounding the radar has been used at times.

Beam Switching for producing Split

In the case of a dipole array on metre wavelengths, the array can be fed in two halves, the phase of the signal in one half being alternately retarded and advanced with respect to the other. This periodic phase change will swing the beam to right and left (or up and down) alternately. The necessary phase change can be produced mechanically or electronically.

At centimetric wavelengths the necessary beam swing can be produced by causing the dipole or waveguide feed to a parabolic aerial to oscillate mechanically over a small distance about its mean position. An alternative method is to mount the dipole assembly slightly off-centre and rotate it about the axis of the mirror. This gives a beam sweeping over the surface of a cone, and provides split for both bearing and elevation, the pairs of signals for the two purposes being separated by a switch ganged to the spindle.

As the dipole must be rotated at a reasonably high speed, this eccentric arrangement is objectionable mechanically. By adjusting the position of the unbalance to balance converter it is possible to make the standing waves of current in the external conductor beyond the point of connection of the dipole (Fig. 10.9) of such magnitude that the radiation from this part, when combined vectorially with that from the dipole, gives a slew to the beam without the necessity of mechanical asymmetry.

Accuracy of Bearing and Elevation Measurements

Without split, angular measurements can be made to an accuracy of about half the beam width. The use of split increases the accuracy about tenfold.

In very accurate work some degree of uncertainty may be introduced if the target is large enough to subtend an appreciable angle at the radar aerial. In the case of a radar used

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for feeding data to a predictor for anti-aircraft fire control (where steady rates of change of bearing are necessary if the predictor is to work accurately), the random wandering of the effective centre of reflection from one end of the aircraft to the other may set a limit to the accuracy obtainable. This random wandering is known as "glint". Fading, and "flutter" due to echoes from the rotating propeller, may also cause difficulty.

Corner Reflector

A reflector consisting of three plane conducting sheets at right angles (Fig. 10.16) will reflect an incident wave back along a path parallel to that by which it came.

A corner reflector, or a cluster of them pointing in different directions, can be used to increase the echo from a buoy at sea or from a meteorological balloon.

The equivalent echoing area of a corner reflector whose sides are triangles of short side a is $4\pi a^4/3\lambda^2$. For example, suppose $a = 50$ centimetres, the equivalent echoing area for a wavelength of 3·2 centimetres (as used in civil marine radar) is 256 square metres.

Lenses

Instead of a parabolic reflector a lens may be used to produce a narrow beam of radiation. A lens of the optical type, made of suitable dielectric, would be too heavy for practical use, but the same effect can be obtained by using a structure consisting of a series of metal plates, as shown in Fig. 10.17 (*a*).

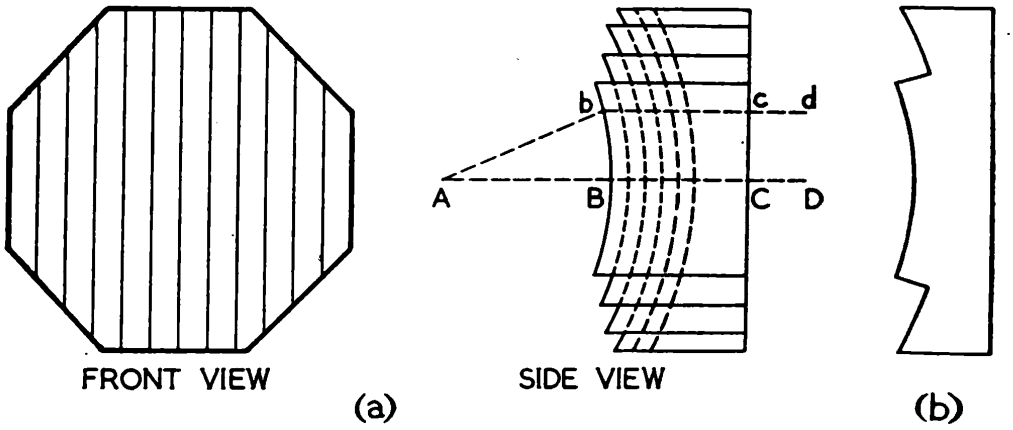

FIG. 10.17.—LENS.

The plates act in a manner analogous to a waveguide, increasing the phase velocity of the wave, with the result that the time taken for the wave to traverse the path *Abcd* is the same as for *ABCD*. A concave lens of this type is equivalent to a convex lens in optics.

To economize in cost and weight of material, the plates may be stepped, as shown in Fig. 10.17 (*b*), the portion removed

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being of the correct dimensions to correspond to a change of a whole number of wavelengths, so that the resultant phase of the emergent beam is unaltered. This stepped construction, by minimizing the length of the path within the lens (where the effective length is sensitive to frequency change), increases the bandwidth of the arrangement.

The metal plates forming the lens must be parallel to the plane of polarization.

Common Aerial Working (Duplexing)

Most radars use the same aerial for transmitting and receiving, the necessary switching being accomplished electronically as shown diagrammatically in Fig. 10.18.

For clarity, the feeders are shown in the diagram as two-wire lines and the switches as spark-gaps. T.R. is the transmit-receive switch, called the T.R. cell, whilst T.B. is the transmitter blocking switch (T.B. cell or A.T.R. cell).

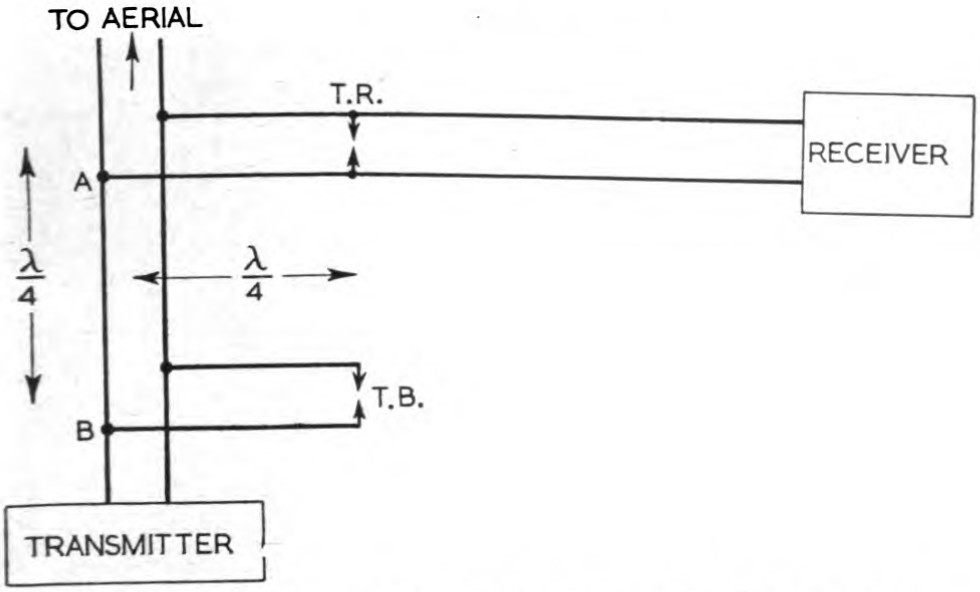

FIG. 10.18.—PRINCIPLE OF T.R. SWITCHING (DUPLEXING).

The action of the circuit depends on the fact that the input impedance of a quarter-wave line is zero or infinite, according as the far end is open or short-circuited.

When the transmitter fires, both spark-gaps break down, forming short-circuits across the lines. T.R. being short-circuited, the shunt impedance of the receiver branch is infinite at the point *A*, and therefore imposes no load across the main feeder, ignoring the trivial amount of power required to maintain the spark. The T.B. cell branch similarly has infinite impedance at *B*.

When the transmitter is not oscillating, the spark-gaps are open and the T.B. branch short-circuits the lines at *B*, resulting

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in infinite impedance at A, which is a quarter-wavelength away. The aerial is therefore connected to the receiver without any loading due to the transmitter.

In some cases the impedance of the transmitter, when not oscillating, is sufficiently different from the feeder impedance to render the loading effect of the transmitter unimportant and the T.B. cell unnecessary.

T.R. Cells

The simple circuit shown in Fig. 10.18, although used for earlier radars on metre wavelengths, must be modified somewhat to suit centimetric conditions. This is mainly because the impedance of the spark-gap when conducting is not zero, and therefore some power would leak into the receiver circuit and burn out the mixer crystal.

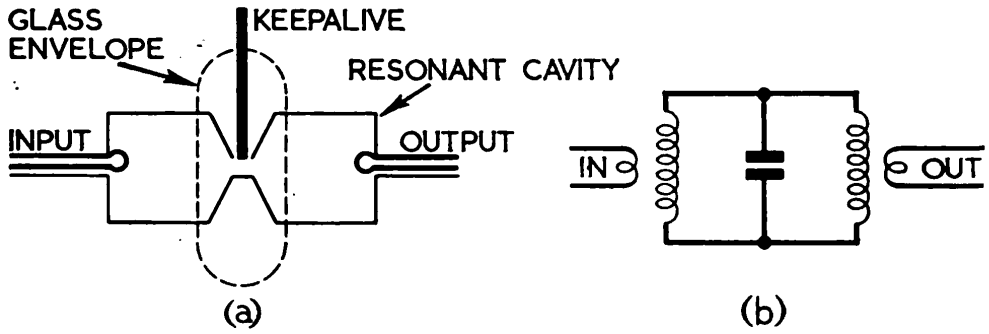

FIG. 10.19.—T.R. CELL.
(*a*) Actual. (*b*) Equivalent circuit.

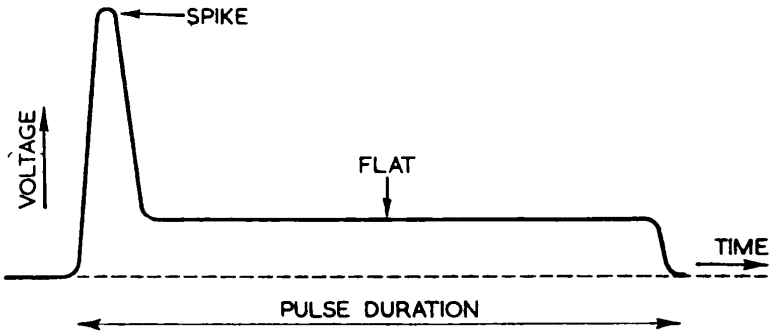

FIG. 10.20.—POWER LEAKAGE THROUGH T.R. CELL.

Practical T.R. cells for centimetric waves are made with resonant cavities [Fig. 10.19 (*a*)] which, by transformer action, raise the voltage across the gap sufficiently to break it down and yet leave only a small voltage across the output leading to the receiver. The equivalent circuit is shown in Fig. 10.19 (*b*).

The cavity is tuned by plungers which can be screwed in, as in the case of the klystron.

By careful design it is possible to keep the loss in the T.R.

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cell on reception down to about 1 decibel and the " breakthrough " power during transmission to about 50 milliwatts.

The gap does not break down instantaneously on the start of the transmitter pulse, and a large amount of power may pass to the receiver before the gas in the cell has become sufficiently ionized for the cell to function. This is illustrated in Fig. 10.20, where the " spike " represents initial breakthrough power, and the " flat " the leakage after the gap has broken down.

The duration of the spike is very short, perhaps only about 0·01 microsecond, and to protect the crystal, the total energy passing through the cell in this time, i.e. the duration of the spike, should not exceed about 1 erg (10^{-7} joule). The provision of an auxiliary electrode, the " keepalive ", connected to

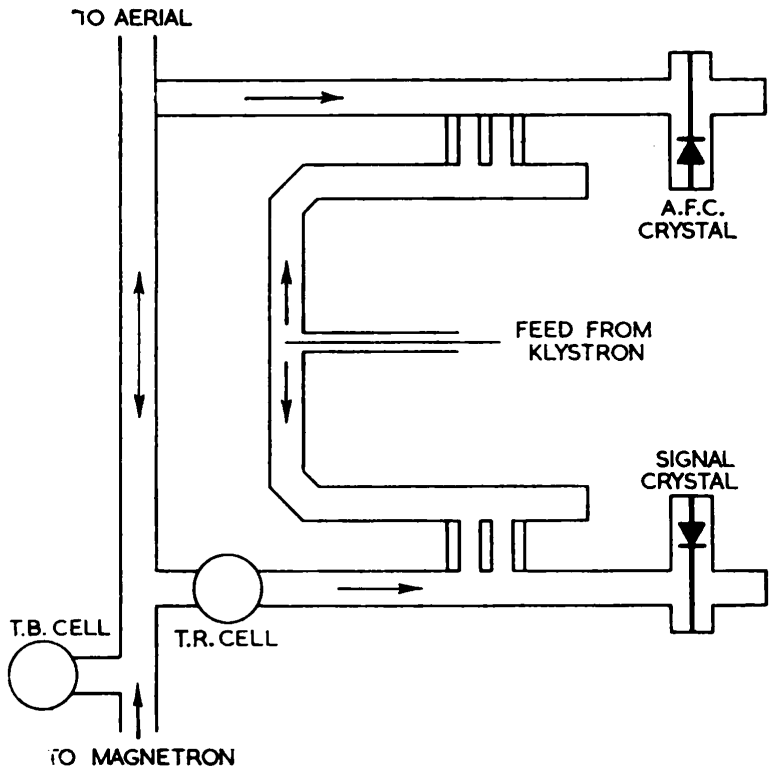

FIG. 10.21.—R.F. HEAD (DUPLEXER) FOR WAVEGUIDE SYSTEM.

a supply at a few hundred volts negative, provides enough ions in the vicinity of the gap to allow it to break down rapidly on arrival of the transmitter pulse. The mean breakthrough during the " flat " should not exceed about 100 milliwatts.

Recovery Time of T.R. Cells

In order that echoes from targets at short range may be received, the cell must recover (i.e. de-ionize) rapidly on the cessation of the transmitter pulse. It has been found that the best gas filling for the cell is water-vapour at a pressure of about 10 millimetres, which gives a recovery time of about

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2 to 3 microseconds, corresponding to ranges of 300 to 500 yards. Echoes from targets at shorter ranges than this can still be received, although at lower strength due to attenuation in the, as yet, partly ionized cell. This attenuation is, however, largely offset by the greater signal strength of echoes from nearby objects.

The addition of a small quantity of hydrogen or argon prevents damage to the crystal should the water-vapour freeze at low temperatures. In these circumstances the recovery time will, however, be longer, but the crystal will be protected.

T.B. Cells

These can be similar to T.R. cells, but the keepalive may be omitted since there is no need for extreme rapidity of action.

Application to Waveguides

The principles are the same as for concentric line feeders, but the cells are coupled to the waveguide by apertures instead of by coupling loops. Fig. 10.21 shows a typical R.F. head or duplexer for a waveguide system.

A more complicated system, employing two ratraces, is shown in Fig. 10.22. This operates as follows. During transmission power cannot pass direct to the aerial, but divides between the two waveguide branches 1 and 2. The T.R. cells break down, forming short-circuits which reflect the power back to the upper ratrace. Owing to the difference in path lengths, due to the T.R. cells being at different relative positions along the two branches, the reflected powers arrive back at the upper ratrace in correct phase for passing to the aerial and not back into the magnetron.

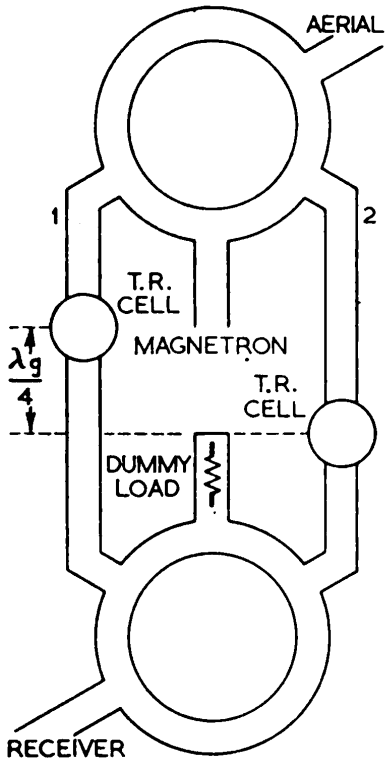

FIG. 10.22.—DUPLEXER EMPLOYING TWO RATRACES.

Breakthrough power from the two cells arrives at the lower ratrace so phased that it all passes to the dummy load and not to the receiver (assuming both lots of breakthrough power are equal, an assumption which is approximately correct). There is thus double protection, and this circuit is suitable for high-power radars.

On reception, power from the aerial passes through the two ratraces to the receiver, none going to either the magnetron or the dummy load.

A similar arrangement can be made by using magic tees instead of ratraces.

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Broadband T.R. Cells

Since a resonant cavity is a sharply tuned device, it is necessary on some high-power radars to make provision for the fact that a magnetron does not emit all its energy on a single frequency, and also the further fact that different magnetrons, although nominally the same, will have slightly different frequencies. In addition, a small amount of power may be generated at frequencies which are harmonics of the fundamental.

To overcome these difficulties, the broadband cell shown diagrammatically in Fig. 10.23 is used. This contains three

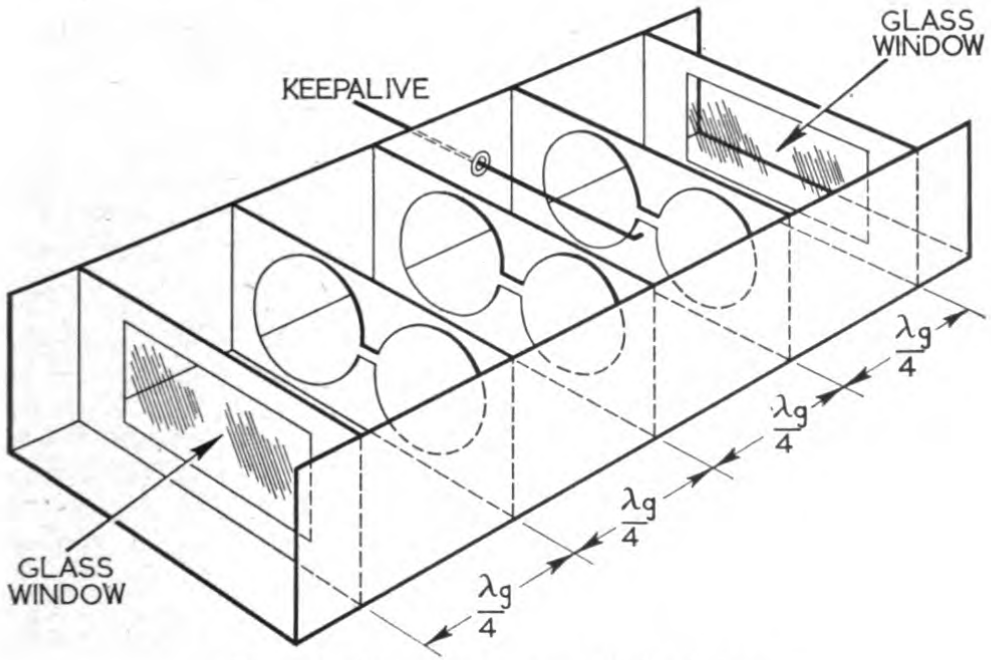

FIG. 10.23.—BROADBAND T.R. CELL.

gaps set in the middle of resonant irises in the waveguide. The tuning of the three irises is staggered, and in any case their Q is not very high, so that a broadband effect is obtained. When working, each iris reduces the leakage power, and the amount finally getting through is very small. The glass windows are to confine the gas filling. A keepalive electrode is fitted close to one or more of the gaps.

This type of cell actually forms a short section of waveguide, and can be removed for replacement when necessary.

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CHAPTER XI

DATA TRANSMISSION AND SERVO SYSTEMS

CERTAIN applications of radar, e.g. fire control and early warning, necessitate the continuous transmission, to a more or less distant point, of information regarding the range, bearing and elevation of a target or targets.

The magslip is the basis of most methods in use.

The Magslip Transmitter

This has a rotor having a single winding, and a stator with three windings spaced 120° apart. If two magslips are connected as shown in Fig. 11.1, and the rotor of *A* is turned to the position shown, voltages will be induced in the stator windings and will produce a resultant field in *B* parallel to the

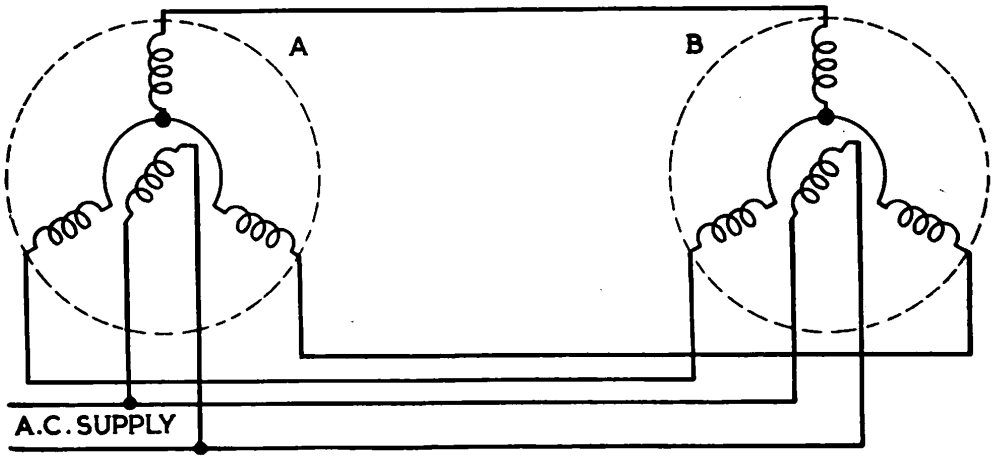

FIG. 11.1.—MAGSLIP TRANSMISSION.

direction of the rotor field in *A*. Rotor *B* will therefore set itself in the same direction.

Angular information, such as bearing or elevation, can be transmitted to *B* by gearing rotor *A* to the bearing or elevation drive of the aerial, as the case may be. Range can also be transmitted by gearing the magslip to a range potentiometer.

The accuracy obtainable is about one degree, but the addition of a second magslip geared up, say 36 to 1, will give correspondingly greater accuracy. In this case, the two sets of magslips are known as low speed and high speed respectively.

The resemblance of the stator winding to that of a three-phase alternator will be noticed, but it should be understood that only a single-phase supply is used. The three windings may be connected in star (as shown) or in delta.

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For some purposes, the stator has only two windings, at right angles, analogous to a two-phase machine.

Magslips are usually designed to operate on 50 to 60 or on 400 to 500 c/s. They are not suitable for transmitting appreciable power.

The Magslip Receiver

Where the receiving magslip is only required to turn a pointer to indicate an angle, it is usual to replace the second magslip (*B*) by a magslip receiver. This has a similar stator winding to the transmitter, but the rotor consists of a suitably shaped member of soft iron with a stationary polarizing winding—see Fig. 11.2. This eliminates the need for sliprings and reduces friction, thus improving the accuracy of indication. The rear end of the spindle bears against a diaphragm which is vibrated by the alternating flux to a small extent, and this still further reduces static friction.

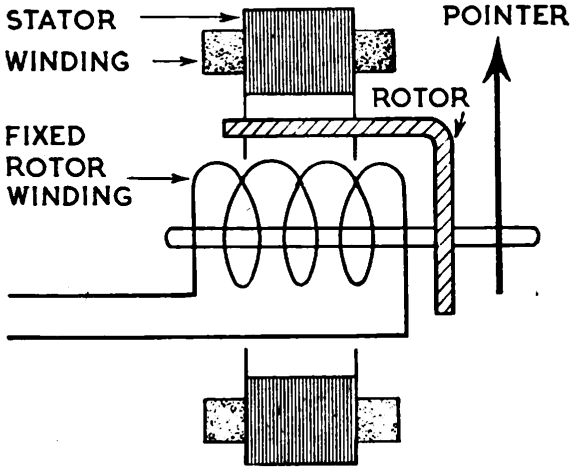

FIG. 11.2.—MAGSLIP RECEIVER.

The Differential Magslip

This has three windings on both stator and rotor, and is used when it is desired that the receiver should indicate the sum or difference of two separate inputs—see Fig. 11.3. If rotor *A* is turned through angle θ from some arbitrary reference point,

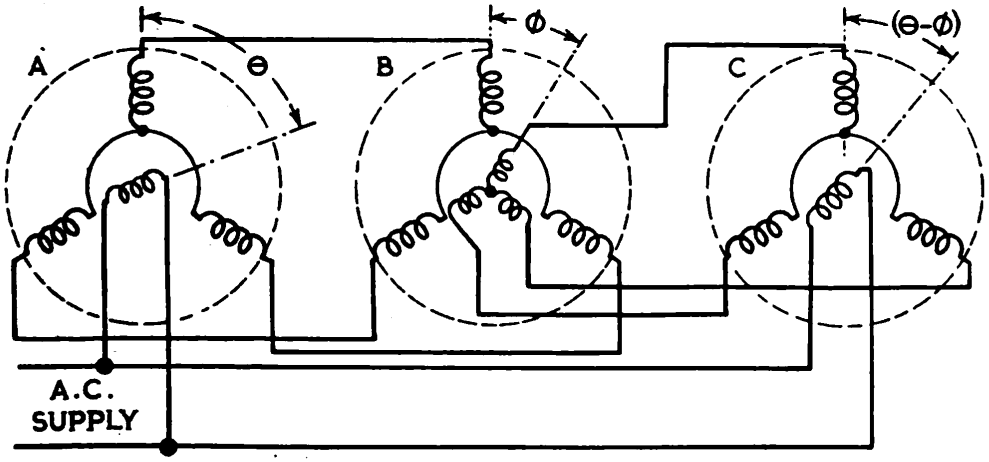

FIG. 11.3.—DIFFERENTIAL MAGSLIP CIRCUIT.

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the field induced in stator B will also be inclined at the same angle. If rotor B is turned through angle ϕ, the direction of the stator field of B relative to the rotor is inclined at angle $\theta - \phi$, and rotor C will take up a corresponding position, as shown.

Sine-cosine Potentiometer

This device, shown diagrammatically in Fig. 11.4, enables two outputs, respectively proportional to the sine and cosine of an angle, to be obtained from a single input. The rotating arm

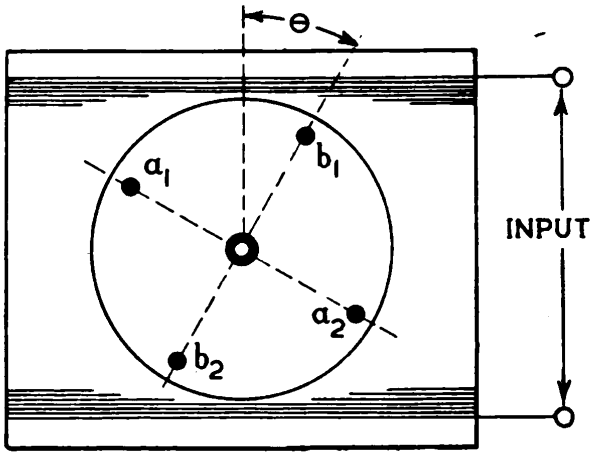

FIG. 11.4.—SINE-COSINE POTENTIOMETER.

carries two pairs of contacts, fixed on lines at right angles to each other.

The sine-cosine potentiometer can be used to drive a magslip-type receiver (having two windings on the stator) or can be used in connection with the production of a rotating timebase for a P.P.I.

Selsyn Low-power Drive

The selsyn is similar to a magslip, but larger, and is capable of handling appreciable power, although the accuracy is somewhat lower due to this fact. Units capable of handling up to several hundred watts can be built, but as there is no power amplification, all the power required to turn the aerial (or other load) at the far end must be supplied mechanically to the rotor at the transmitting end. The rotor can be turned continuously or intermittently as required. It is usual to employ gearing, partly to secure increased accuracy and partly because the speed at which the driven member (aerial, etc.) is to be turned is ordinarily much lower than the speed at which rotating electrical machinery functions efficiently. When gearing is employed, the ambiguity resulting from the fact that one position of the selsyn rotor corresponds to several possible aerial positions must be eliminated by additional means of indication.

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Servo Selsyn

This has a ring-wound rotor and commutator for the input a.c. and three sliprings for the output. See Fig. 11.5. This

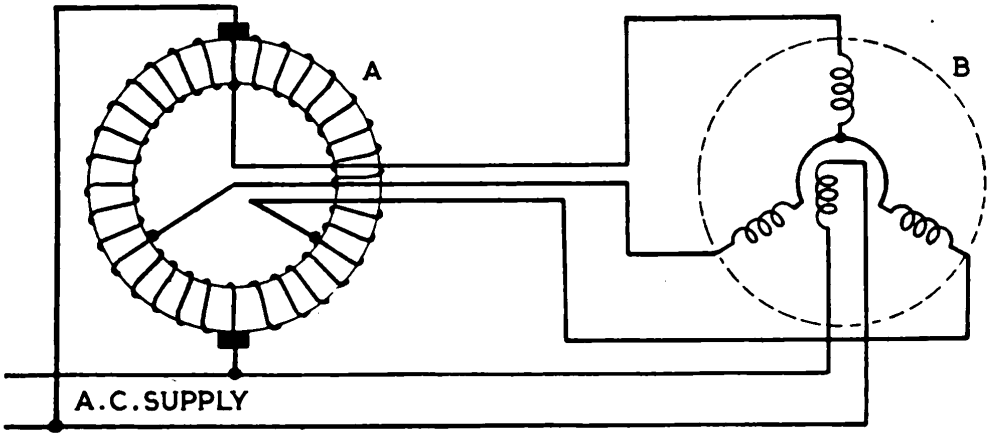

FIG. 11.5.—SERVO SELSYN.

device acts as a power amplifier, the only power required in turning the rotor being that due to friction. If a stator winding is added, the device can be self-driven when desired, e.g. for sweeping.

P.P.I. Applications

With P.P.I. presentation, it is necessary to rotate the direction of the radial timebase on the cathode-ray tube in step with the rotation of the aerial. This can be done by rotating the deflecting coils on the cathode-ray tube by a magslip or selsyn controlled by the aerial position, or by driving both aerial and coils by synchronous motors operating from the same a.c.supply.

A differential gear can be incorporated in the assembly to permit of manual adjustment to bring true north to any desired position on the screen.

An alternative to rotating the deflecting coils on the c.r.t. is to employ a magslip having two stator windings connected to corresponding coils fixed at right angles on the c.r.t. The timebase waveform is passed through the rotor of the magslip, which is mechanically coupled to the aerial. A sine-cosine potentiometer can be used instead of the transmitting magslip.

The introduction of a differential magslip, controlled by gyro compass, in the circuit between aerial and c.r.t. provides a means of stabilizing the direction of the picture with respect to true north in the case of a ship's installation.

Accurate Power Drives. Servo Systems

A radar aerial is usually of appreciable size and weight, and is required to be moved in bearing (or elevation) according to some specified requirement with a high degree of accuracy, and this accuracy must not be adversely affected by inertia, windage,

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etc. This implies the use of a driving motor of considerable power, combined with means for accurately controlling its rotation.

This introduces the conception of an "error signal". An error signal (which may be a voltage, an angular displacement or anything else suitable) is a signal whose magnitude is approximately proportional to the extent to which the actual position of the aerial differs from its required position, and whose polarity or phase depends on the sign of the error.

A servo system is one in which this error signal is suitably amplified and used to control the rotation of the driving motor.

The manner in which the error signal is produced depends on the nature of the installation. For example, if the aerial is required to follow accurately the motion of an operator's handwheel, a magslip transmitter may have its rotor geared to the aerial. A second magslip, now called a resolver magslip, has its rotor geared to the handwheel so that when the aerial

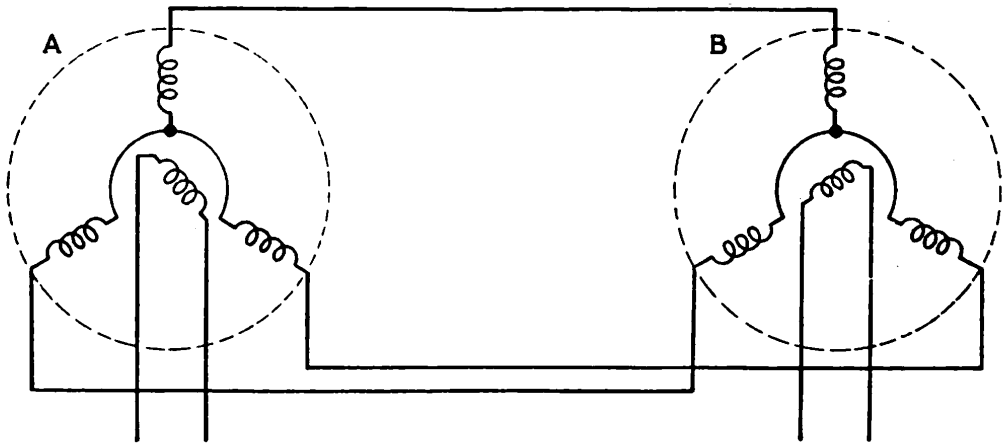

FIG. 11.6.—MAGSLIPS USED FOR SERVO CONTROL.

is in the correct position, the two rotors are at right angles, as in Fig. 11.6. In this condition no voltage will be induced in the rotor of the second magslip. If an error exists in the position of the aerial, the two rotors will no longer be at right angles, and an error voltage will be induced in the second. The magnitude will be proportional to the sine of the angle of error, the phase depending on the sign of the error. This error signal is amplified and used to control the main driving motor which then rotates until the error is reduced to zero (or substantially zero).

The error signal is produced in different ways in different equipments; for example, with auto following (see below) it is produced electronically from the echo signals.

Instead of a resolver magslip, a hunter magslip is sometimes used. This is wound like a differential magslip, the rotor carrying a contact arm moving between two fixed contacts. When the sending and receiving magslips are in alignment, the contact

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arm is held by light springs in a mid-position between the two fixed contacts. Deviation from correct alignment causes one or other contact to be made. Relays connected to these contacts control the driving motor.

Low-power Drive using Two-phase Induction Motor

One phase of the motor (Fig. 11.7) is fed direct from the a.c. supply, and the other from the output, suitably amplified, of a resolver magslip controlled by a transmitting magslip geared to the member it is desired to follow.

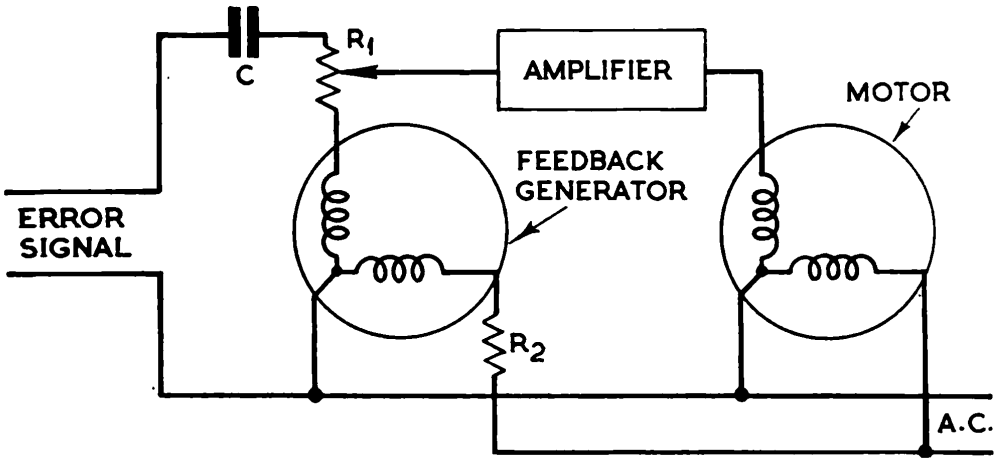

FIG. 11.7.—LOW-POWER DRIVE USING 2-PHASE INDUCTION MOTOR.

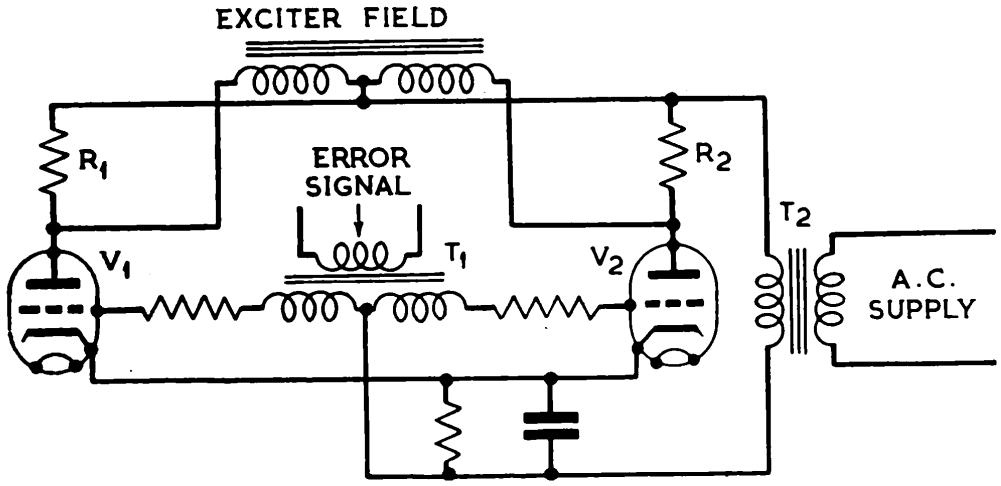

FIG. 11.8.—ERROR SIGNAL AMPLIFIER.

Coupled to the motor is a similar machine which acts as an induction generator, producing a voltage proportional to speed, which is fed back in opposition to the input from the resolver, to provide negative feedback to prevent hunting.

The necessary phase shift between the two supplies to the motor is obtained partly by the capacitor shown, and partly in the amplifier.

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This arrangement is suitable for low-power drives, such as rotating the coils of a P.P.I. tube.

Amplification of Error Signal

In many cases, the initial error signal requires amplification and rectification before it can be applied to the various devices described. This can be done electronically, and a typical circuit is shown in Fig. 11.8. The h.t. supply to the anodes of V_1 and V_2 is raw a.c. from the same supply as that used for the magslips. The error signal from a resolver magslip is applied to the grids of the two valves in opposite phase, and the anode current passed by the two valves is therefore equal when there is no error signal. When an error exists, the voltage applied to the grids helps one and opposes the other, causing the currents in R_1 and R_2 to differ, and consequently varying the primary excitation of the amplidyne or other device used. This arrangement takes care of the phase of the error signal.

Driving Gear

For low-power work it is possible to amplify and rectify the error signal and use it to excite the field of a small d.c. motor whose armature is supplied with a substantially constant current. Provided the amplifier has phase-discriminating properties, the direction of rotation of the motor will correspond with the sign of the error signal.

Usually, however, greater power amplification is required, and typical examples of commonly used arrangements are the metadyne and the Ward-Leonard systems.

Metadyne and Amplidyne

Fig. 11.9 shows the principle of operation of the metadyne. Basically this consists of a constant-speed motor M_1 driving a d.c. generator having an additional set of brushes (3 and 4) in addition to the normal two (1 and 2). The latter brushes are short-circuited, and full armature current therefore flows with a very small main field excitation in winding F_1. Armature reaction produces the usual cross-flux at right angles to the primary field, and this cross-flux causes a voltage to be generated across brushes 3 and 4. A compensating winding F_2 carries the main output current and opposes the armature reaction due to this current, which would otherwise swamp the relatively weak primary excitation in F_1. The amount of compensation can be adjusted to suit requirements; when complete compensation is used, the device is known as an amplidyne.

The metadyne is a two-stage electrodynamic amplifier, the first stage of amplification being the production, from the primary excitation, of the cross-flux due to current in the short-circuited brush circuit, and the second stage the generation of power across the other brushes from this armature reaction, now used as the main field. Power amplification up to about 20,000 can

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be obtained. The actual power is, of course, provided by the motor driving the metadyne.

The primary excitation is provided by the amplified error signal, and its direction depends on the sign of the error.

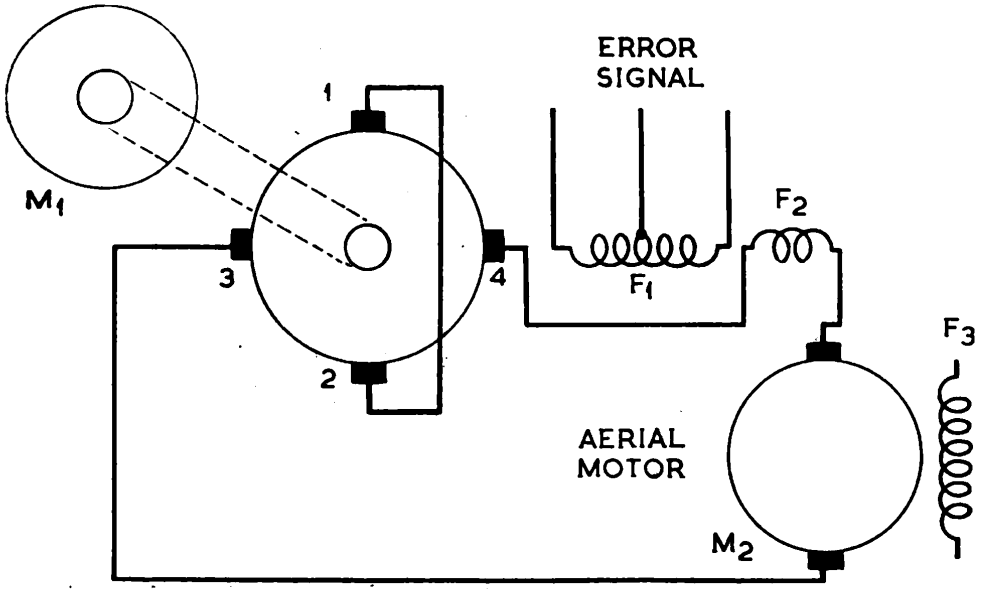

FIG. 11.9.—METADYNE DRIVE.

As the amplification of the device is very great, it is necessary to prevent the production of appreciable output voltage due to remanent magnetism when the primary excitation is zero. This can be done by providing an a.c. demagnetizing field from a small alternator carried on the main shaft of the machine.

Ward-Leonard Control

In this system, the error signal is amplified sufficiently to provide the main excitation for a d.c. generator driven by a constant-speed motor. The magnitude and polarity of the voltage of this generator depend on the error signal, and this variable voltage is used to drive the aerial turning motor, which is separately excited. This arrangement is equivalent to a single-stage power amplifier, and can give a power gain of the order of 100. Two such arrangements in cascade can give a gain comparable with that obtained with the metadyne.

Hunting of Servo Systems

As a servo system involves elements having inertia, and forces more or less proportional to the displacement of those elements from their correct positions, there is an inherent tendency for the system to oscillate mechanically about its mean position at any instant.

By analogy with an electrical circuit containing inductance and capacitance, it would seem that the cure for this might lie

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in increasing friction (the mechanical equivalent of electrical resistance). The deliberate increase of friction is, however, very wasteful of power, and is liable to lead to inconsistency in action.

Waste of power can be avoided by providing a heavy flywheel connected to the driven member by a friction drive. This provides the desired friction when acceleration occurs, but at other times the flywheel rotates with the driven member without slipping.

This method has been adopted with some success on low-power drives for data transmission, but the flywheel would be impracticably large if used with a heavy aerial.

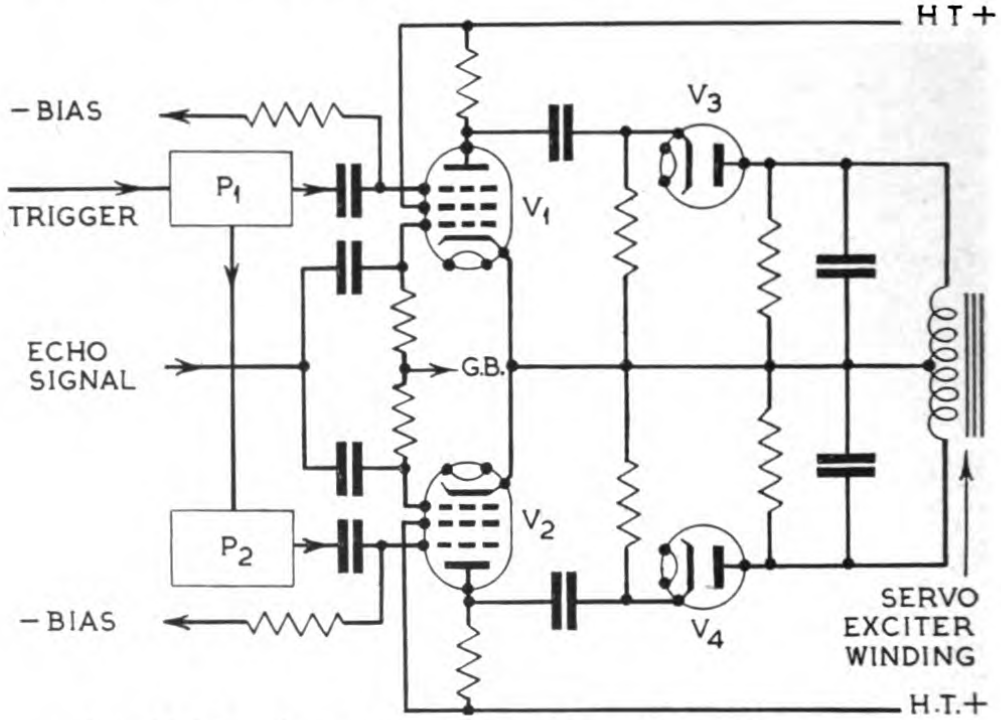

FIG. 11.10.—AUTOMATIC FOLLOWING—BASIC CIRCUIT.

Another method is to connect a tachometer (a small d.c. permanent-magnet generator giving an output voltage proportional to speed) to the driven member and feed the output of this to the control mechanism in opposition to the error signal. This is an electrical method of producing a retarding force proportional to speed.

The foregoing expedients are, however, only partial solutions. A servo system is a power source and can, in certain circumstances, hunt even if the precautions detailed above are taken. A simple example will suffice. Suppose there is a time lag in the response of the servo system. When the driven member reaches its correct position, it *must* overshoot it, because a short time will elapse before it "knows it has got there". This overshoot is then corrected, and followed by one in the opposite direction, i.e. the system will continue to hunt.

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This can be dealt with by using phase-shifting circuits which give a lead to the error signal at the frequency at which the system tends to oscillate, this lead cancelling the lag referred to above.

The problem is considerably more complicated than the short statement made here would imply, but space does not permit of fuller treatment.

The difficulties mentioned and the steps that must be taken to overcome them prevent a servo system from being absolutely accurate. For example, assuming steady rotation of an aerial, there will always be a small permanent positional error, and it is the aim of the designer to make this as small as possible.

Automatic Following

Fig. 11.10 shows the elements of a circuit that will enable a radar to follow automatically in range a selected target once it has been "put on" to that target by the operator.

Video echo signals are fed in parallel to two pentodes V_1 and V_2. P_1 and P_2 represent two gate-pulse-forming circuits which generate gate pulses, of length comparable with that of the echo. P_1 is triggered by a pulse from a range potentiometer, and P_2 is triggered by the trailing edge of the pulse from P_1. The result is two pulses displaced in time as shown in Fig. 11.11.

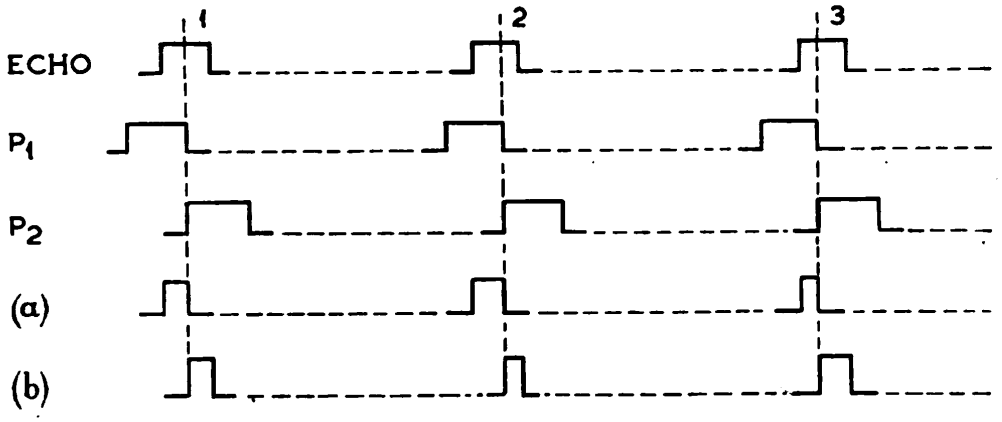

FIG. 11.11.—AUTOMATIC FOLLOWING—GATE PULSES.

V_1 and V_2 are biased so that each conducts only when there is both a gate pulse and an echo pulse.

The outputs of V_1 and V_2 are passed through the diodes V_3 and V_4, and their difference is the error signal. This is applied to the field winding of a servo motor controlling the setting of the range potentiometer.

If the two gate pulses, whose position depends on the setting of the range potentiometer, are symmetrically disposed with respect to the echo signal, the error signal is zero; otherwise it is in a direction which alters the setting of the range potentiometer until the error is reduced substantially to zero.

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Fig. 11.11 shows three cases corresponding respectively to zero error, echo early and echo late.

A magslip, geared to the slider of the range potentiometer, conveys the range information to the position at which it is required. As the first gate pulse must start before the instant at which the echo arrives, a " zero correction " must be made when setting up the slider or magslip initially.

Circuits for producing trigger and gate pulses are described in Chapter XV.

For following in bearing or elevation, an aerial system using split is used, the echo signals corresponding to the two beam positions being fed through gated circuits of the same general type as Fig. 11.10. The servo system then drives the aerial round until the outputs from the two halves of the circuit are equal. To avoid difficulty due to two targets being on the same bearing but at different ranges, it is desirable to provide an overriding gate circuit, controlled by the ranging circuit, which only allows echo signals from a target at the correct range to reach the auto-follow circuit.

The time constants of the output circuit can be made long enough to avoid losing the target during periods of momentary fading, but there must be a compromise between this facility and the required rapidity of response when the rate of change of range or bearing alters. It is also useful to provide the operator with a switch for cutting out the incoming signals when the visual display suggests that a stronger unwanted echo may take charge, or heavy clutter cause the wanted echo to be lost. Whilst this switch is open, the circuit carries on as during fading.

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CHAPTER XII

CATHODE-RAY TUBES

CATHODE-RAY tubes are very widely used for radar display purposes because of their versatility and extreme rapidity of action.

There are two main types, employing electrostatic and magnetic deflection, respectively, of the electron beam. Either type may be intensity modulated by control of the voltage on the grid relative to the cathode. If the grid is made sufficiently negative, the beam is cut off; as it is made less negative, the spot appears, and its brightness increases as the negative voltage is further decreased.

Construction

The construction of the electrostatic tube is shown in Fig. 12.1. The cathode is a cylinder, indirectly heated, and the "grid", which is also cylindrical, has a small hole through which the electron stream emerges. The intensity of this stream depends on the grid voltage relative to the cathode. The remaining anodes are also cylindrical.

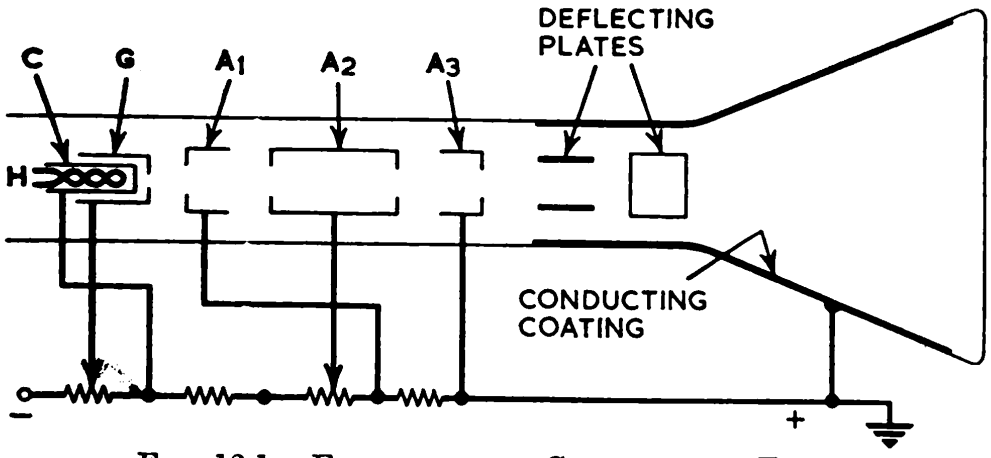

FIG. 12.1.—ELECTROSTATIC CATHODE-RAY TUBE.

The beam is deflected as required by the application of a voltage between the appropriate pair of deflector plates. There are two pairs of plates, for horizontal and vertical deflection, respectively.

The principle of the magnetic tube is shown in Fig. 12.2. The cathode and grid system is generally similar to that of Fig. 12.1, but deflection is by the application of a magnetic field perpendicular to the axis of the tube.

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Focusing

The electron stream emerging from the hole in the grid is divergent, and must be focused so as to produce as small as possible a spot at the screen.

Electrostatic tubes are generally electrostatically focused, whilst magnetic focusing is employed for magnetically deflected tubes. There is no fundamental reason, however, why magnetic focusing should not be employed with electrostatically deflected tubes, and vice versa, and in fact this is sometimes done for special purposes.

With electrostatic focusing, the electron beam, having left the grid (Fig. 12.1), is accelerated by the first anode A_1 and then passes into the field of the second anode A_2 which is at

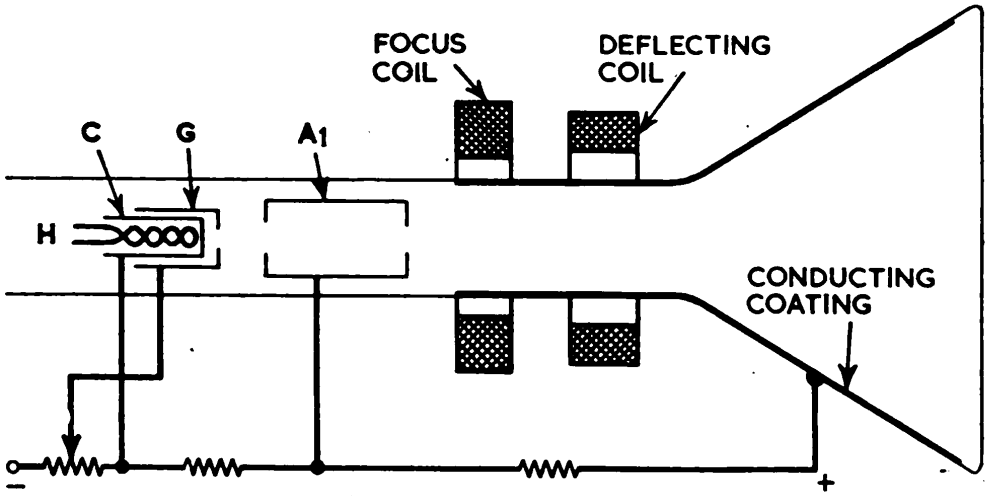

FIG. 12.2.—MAGNETIC CATHODE-RAY TUBE.

a lower (adjustable) potential. This repels the electrons, causing a " compression " of the beam so that, by the time the screen is reached, the beam has become practically a line of very small diameter. The third anode A_3 provides further acceleration.

In the magnetically focused tube (Fig. 12.2) a coil around the neck of the tube produces an axial magnetic field. Since electrons diverging from the centre of the beam have a radial component of velocity, this axial field produces a circular motion of the electron. The coil is so designed that the circular motion imparted to the electron by the time it leaves the influence of the focusing field is of the correct magnitude and direction to bring the electron to the centre line by the time the main longitudinal velocity has brought it to the screen.

Brightness

The brightness of the trace depends on the grid voltage, i.e. on the density of the electron stream, and also on the final anode voltage, which determines the final velocity of the electrons.

Brightness can be increased by increasing the final anode voltage, but this reduces the deflection sensitivity. In some

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cases tubes have an additional electrode, at a higher voltage, for further accelerating the electrons after deflection (called post-deflection acceleration).

Deflection Sensitivity

In an electrostatic tube, a potential of v volts applied between a pair of deflecting plates will produce a deflection at the screen of $\frac{adv}{2bV}$, where d is the distance from the middle of the plates to the screen, a is the axial length of the plates and b their distance apart. V is the final anode voltage, with respect to the cathode. In practice, sensitivities of from 0·1 to 0·5 millimetre per volt are common.

In the magnetic tube, the deflection is $adB\sqrt{\frac{e}{m}\cdot\frac{1}{2V}}$, where B is the flux density of the transverse deflecting field, a is the axial length over which it acts, and d and V are screen distance and final voltage as before. Inserting numerical values for e and m, the formula becomes $\frac{adB}{3{\cdot}37\sqrt{V}}$, the units being gauss, centimetres and volts.

Usually two coils, at right angles, are used for horizontal and vertical deflection respectively. In the case of a P.P.I. display where the radial timebase is required to be rotated continuously, a single coil, arranged for mechanical rotation about the neck of the tube, is often used. Sometimes, however, it is preferred to use two coils and apportion the current between them to give the required rotational effect (see page 138).

Relative Merits of Electrostatic and Magnetic Tubes

The electrostatic tube is preferable for displays of the A-type because the rapidly varying voltages required for deflecting the beam are more easily generated than the corresponding currents which would be required for the magnetic tube. In the case of a P.P.I., however, the echo signals are used to cause intensity modulation of the beam, and the deflecting currents required to produce the timebase trace vary relatively slowly. These facts make it better to use the magnetic tube, with its superior focusing properties, for this display.

Astigmatism

In an electrostatic tube, if the mean of the voltages on one pair of deflecting plates, relative to the final anode, is incorrect, there may be defocusing of the spot in the horizontal plane and not in the vertical, or vice versa. This is corrected by applying a suitable (adjustable) voltage between one pair of plates and the final anode.

Similar defocusing may occur if the deflecting voltage is

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applied to only one plate of a pair, the other being at constant potential. For this reason, a push-pull circuit, applying equal and opposite deflecting voltages to the two plates, is preferred.

Centring of Spot

In an electrostatic tube this is done by applying small d.c. voltages to one or both pairs of deflecting plates as required. In the magnetic tube, a pair of auxiliary coils at right angles, fed with d.c. from separate potentiometers, performs a similar function.

An extension of this principle permits the whole display to be offset for particular purposes.

It is important to shield the tube from external magnetic fields by means of a mumetal shield.

Ion Burn

As perfect evacuation is impossible, a number of ions will be present in the tube. Owing to their greater mass, these ions will move with much lower velocities than the electrons. In the electrostatic tube they will consequently remain under the action of the deflecting field for a longer time, and this will compensate for their smaller acceleration due to their larger mass. The deflection of the ions will be the same as the electrons.

With the magnetic tube, however, since the deflecting force depends, among other things, on the velocity of the particle concerned, ions being relatively heavy will hardly be deflected at all. All the ions will therefore strike the screen at or very near the centre, causing, after a time, burning of the screen.

Tubes can be constructed to overcome this by having their electron guns bent so that, initially, electrons and ions are both directed at a point off the screen. A small permanent magnetic field is added to deflect the electrons in the required direction, leaving the ions trapped in the gun assembly.

Screens

For displays where the trace is repeated at short intervals (many times per second), as in A-type displays, the persistence of vision makes "afterglow" properties unnecessary, and the chief consideration is high visual efficiency. For such purposes screens made of willemite (zinc orthosilicate) or certain sulphides are suitable.

In the case of displays, such as the P.P.I., where the trace over any particular part of the screen is only repeated at intervals of some seconds, a long afterglow is required.

Magnesium fluoride screens giving an orange phosphorescence are suitable for this purpose, and are being increasingly used. Formerly it was the practice to use "double layer" screens, the first screen, having no afterglow properties, being excited by the electrons, and the light from this being used to excite

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the second screen, which was made of suitable material to provide afterglow.

The rear surface of the screen is frequently sprayed with finely divided aluminium which increases the brightness by reflecting some of the light which would otherwise pass back into the tube and be wasted. This aluminium coating offers negligible resistance to the passage of the electrons to the screen.

Screen diameters in common use vary from about 2 to 15 inches.

Electrode Potentials

Radar tubes are worked at voltages (relative to cathode) of from 2 to 15 kilovolts on the final anode. The grid is negative to cathode, a voltage of 50 to 100 giving complete cut-off, and lower values varying degrees of brightness. If an electrostatic tube is provided with three anodes, the first will usually be at a potential of about 40 per cent of the final, whilst the potential of the second (the focusing anode) will be adjustable between limits of roughly 10 to 20 per cent of final anode voltage.

The required voltages can be derived from a potential dividing chain as shown in Figs. 12.1 and 12.2. Currents are negligible.

Heaters require about one or two watts, usually at 6·3 volts.

Picture Storage Tube

This has two electron guns and a screen of metal covered with a thin film of insulating material. The electron beam from one gun, known as the writing beam, scans the screen and creates on it a charge of density varying from point to point in accordance with the varying intensity of the beam.

The beam from the second gun, called the reading beam, is of low intensity and also scans the screen, usually at a much higher rate, and discharges each point by a small amount per scan proportional to the charge at each point.

These operations are made possible by the fact that the secondary emission ratio of an insulator (the screen) under electron bombardment is less for higher electron velocities than for low. The voltages of the two beams are adjusted so that the writing beam operates at a voltage corresponding to a secondary emission ratio less than unity, i.e. it causes a net gain of electrons on the screen. The reverse is the case with the reading beam.

The reading beam signals are taken from the metal backplate, amplified, and displayed by television methods.

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CHAPTER XIII

RADAR DISPLAY SYSTEMS

THE type of display used in any particular installation depends on the purpose for which it is intended. Requirements may vary from accurate range bearing and elevation of a particular target to a general indication of all targets within range.

The commoner types of display are shown in Fig. 13.1.

A-Type Display

This, shown at (i), indicates range on an electrostatically deflected cathode-ray tube. The spot is deflected from left to right at a uniform rate (controlled by the timebase generating circuit), and echoes are indicated by vertical deflections produced by applying the amplified signals to the *Y*-plates of the tube. Range is indicated by the distance along the trace at which the echo appears.

Since the whole range is comprised within the diameter of the screen, high accuracy is not obtainable.

For greater accuracy the timebase trace may be considerably extended, and only a small part, containing the echo concerned, shown on the tube, as in (ii).

Another method is to speed up the part of the timebase which contains the wanted echo, as in (iii).

Where one operator has to indicate to another that a particular target is to be dealt with, he can do this by applying a " strobe " which causes a brightening of the trace over a small portion containing the echo, as in (iv), or by moving a step along the trace, as in (v).

The circuits for producing the above effects are described under timebases.

The foregoing displays take no account of bearing or elevation, these being obtained from the angular positions of the aerial in the horizontal and vertical directions. When split is employed, the echoes corresponding to the two positions of the beam may be displayed side by side, slightly displaced from one another, as in (vi), to assist the operator in turning the aerial to obtain equality of signal strength from the two positions. The small displacement required can be obtained by applying a small voltage in series with the timebase voltage for one position of the beam. Switching may be electronic or mechanical, according to the method adopted for swinging the beam.

B-Type Display

This shows range plotted against bearing, as in (vii). The horizontal deflection is produced by a voltage derived from a

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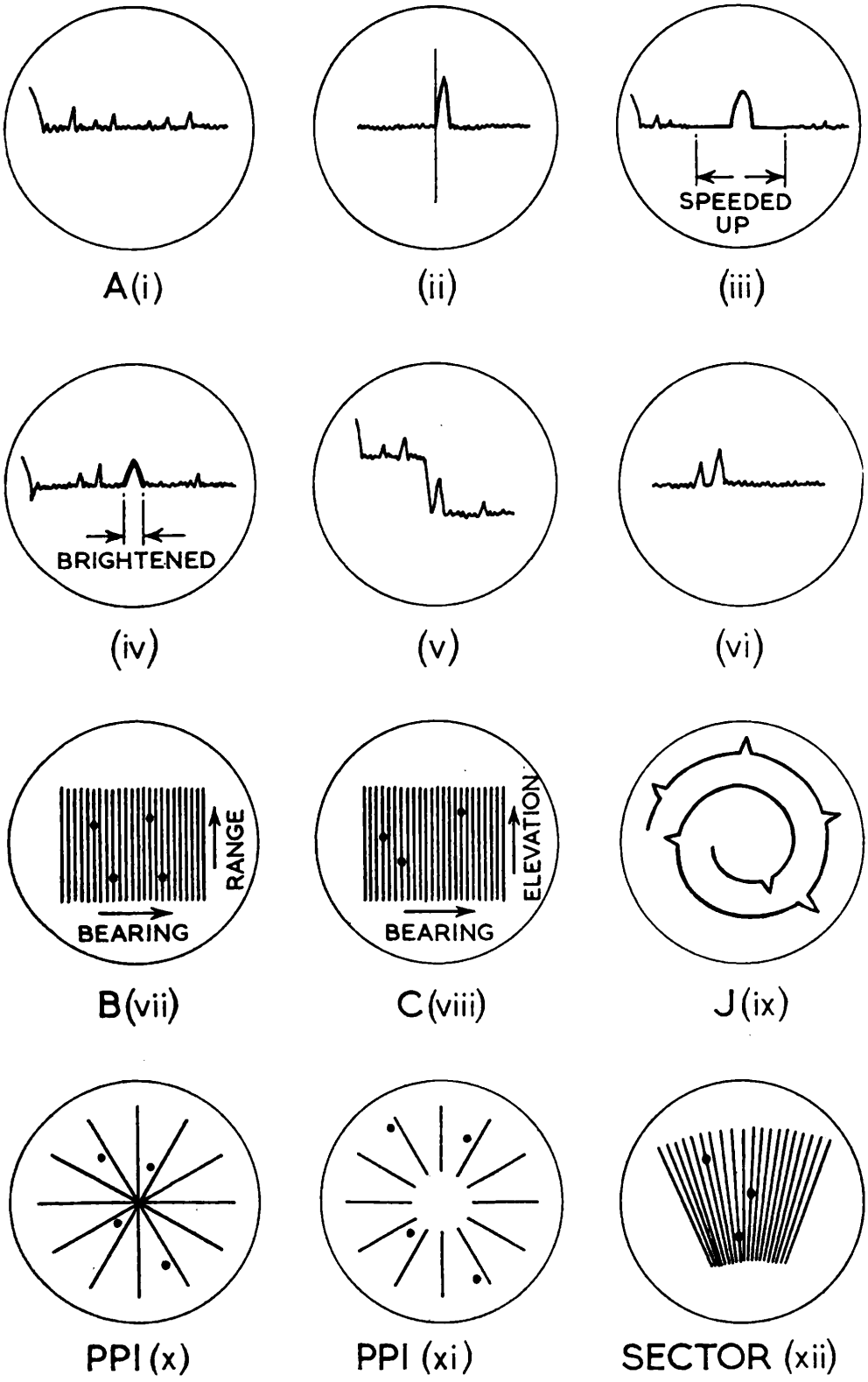

FIG. 13.1.—TYPES OF DISPLAY.

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potentiometer mechanically coupled to the aerial, the vertical range sweep is obtained from a linear timebase generator, whilst the target is indicated by a momentary brightening of the spot.

C-Type Display

This, shown in (viii), is similar to the B-type, except that range is replaced by elevation. The elevation deflection can be obtained from a potentiometer coupled to the aerial.

This type of display has been used on fighter aircraft to indicate when an enemy aircraft was dead ahead.

J-Type Display

This, shown at (ix), is similar to the A-type, but has a spiral instead of a straight-line timebase, thus giving greater length with correspondingly improved accuracy.

Echo signals are shown by outward radial deflections produced by applying the signal to a special electrode at the centre of the screen. A graduated scale and a cursor pivoted at the centre of the screen afford a means of reading off the range of any echo.

P.P.I. Display

The "*P*lan *P*osition *I*ndicator" display, shown at (x), has a radial timebase trace starting from the centre of the screen. The direction of the sweep is continuously rotated in synchronism with the aerial, and targets are indicated by a momentary brightening. This display therefore gives a radar map of all targets within range, and is largely used for marine radar and for many types of airborne equipment.

A modification is the expanded centre display shown in (xi), in which the timebase starts, not from the centre of the screen, but from a point at some distance from it. This gives greater discrimination where there are several targets at short range, as may occur with marine installations. The picture is, of course, somewhat distorted because the centre, which should be a point, is expanded into a circle.

With P.P.I. displays, the aerial is normally rotated continuously, at from 6 to 30 r.p.m.

Range is determined by aid of calibration rings produced by brightening the trace at intervals corresponding to certain definite ranges. For more accurate measurements a variable range ring, whose radius is controlled by an accurately calibrated potentiometer, can be adjusted until it intersects the echo concerned; the range is then read off from a calibrated scale attached to the potentiometer.

By altering the speed of the radial timebase sweep different ranges may be obtained.

In the case of a marine installation, provision is often made for automatic stabilization of the picture by the ship's gyro compass, as described in Chapter XI.

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Accuracy obtainable with P.P.I. Display

Bearings can be determined within one or two degrees when the echo appears near the edge of the screen. The actual measurement is made by aid of a cursor, pivoted at the centre of the screen, and an annular degree scale.

Range can be measured to an accuracy of about 2 to 5 per cent of the maximum of the range scale in use. The discrimination possible between two targets close together depends on actual range, pulse length, and other considerations, but at short range it is generally possible to separate two objects if they are from 25 to 100 yards apart.

Modification for Special Purposes

If only targets lying within certain limits of range and bearing are required to be shown on the screen, the starting-point of the timebase sweep may be offset by applying a fixed deflecting field, so that in effect the sweep starts from a point some distance outside the screen. A larger scale may therefore be used over the wanted area. The result is a sector display, as shown in (xii). The limiting case of this is the B-type display.

Large-scale P.P.I. Displays

The production of a large-scale display, suitable for simultaneous viewing by several observers, cannot be achieved by normal optical projection methods because the brightness of the usual afterglow tube cannot be increased sufficiently without making the initial brightness (called " excitation flash ") objectionably great.

The difficulty can be overcome by using a tube without afterglow properties; this can be brightened sufficiently for projection on to a screen coated with a suitable afterglow material, such as zinc-cadmium sulphide.

An exception to this statement is the " skiatron ", which is a tube having a screen coated with potassium chloride, and which gives a dark trace sufficiently pronounced for projection. The usefulness of the method is limited, however, by the excessively long time taken for the echo signal " paints " to decay.

Another solution is to use the storage tube (see Chapter XII), but the added circuitry represents an appreciable complication.

Moving Target Indication

If the echo signals produced by one transmitter pulse are subtracted from those produced by the next, the resultant will contain only signals from objects which have moved between the two pulses; echoes from fixed objects, being unchanged, will cancel out. By this means moving objects, such as aircraft, can often be shown on a P.P.I. in circumstances when they would otherwise be obscured by heavy clutter from nearby fixed objects.

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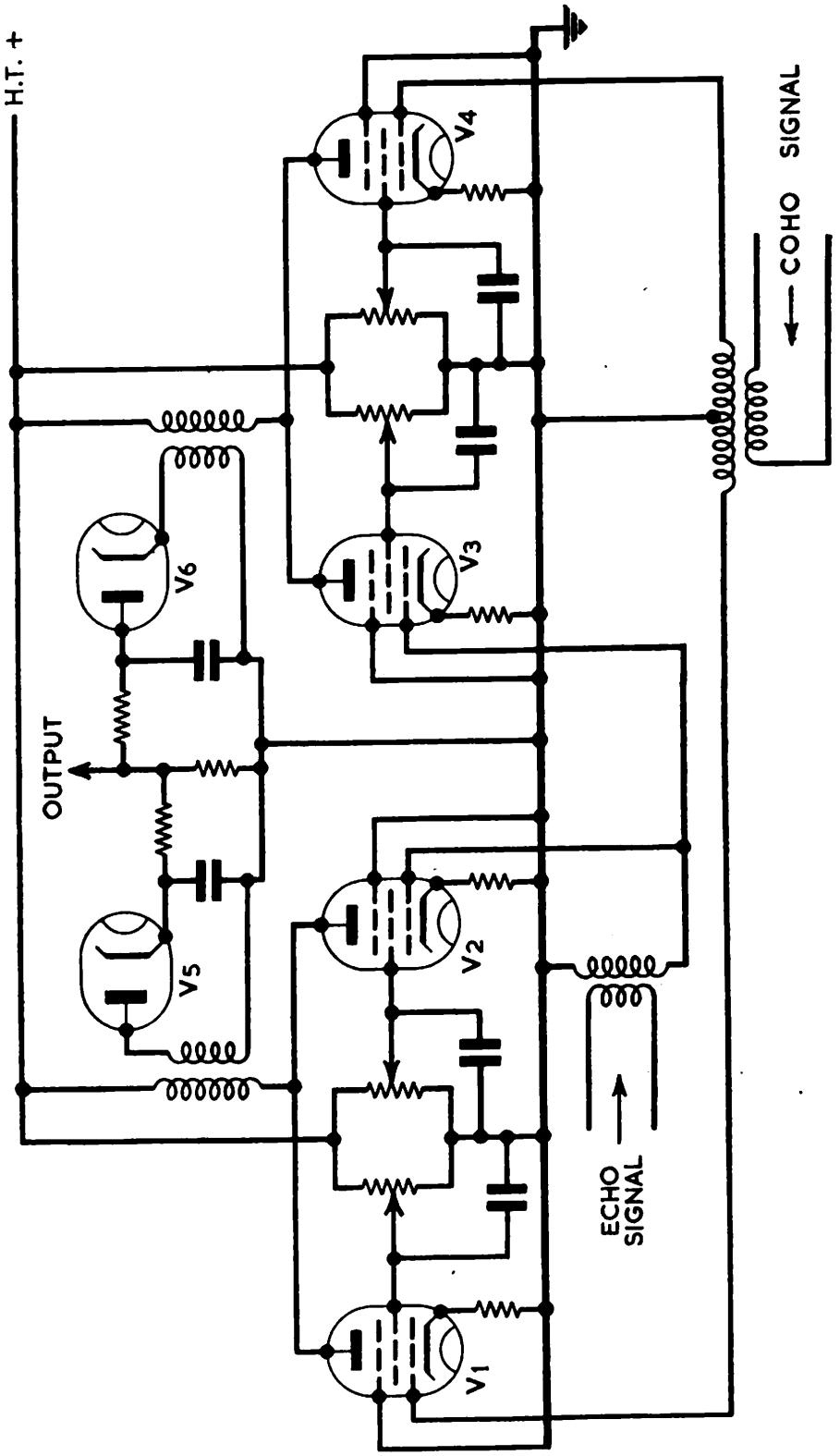

FIG. 13.2.—MOVING TARGET INDICATION—BASIC CIRCUIT.

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In practice the problem is complicated by the fact that the intervals between successive transmitter pulses cannot be relied upon to be constant to the very close degree required for accurate comparison.

This difficulty can be overcome by comparing the phase of an echo with that of the transmitter pulse producing it, and then comparing this phase difference with the corresponding quantity for the next transmitter pulse.

As the transmitter ceases to oscillate before the echo arrives back, the expedient of triggering a stable oscillator (coherent oscillator—" coho " for short) by the transmitter is adopted. This ensures that a correctly phased oscillation is available for comparison with the echo signal.

The coho consists of an oscillatory circuit with associated pentode, arranged to be cut off on its suppressor initially. The transmitter pulse shock excites this oscillatory circuit, and a switching wave shortly afterwards removes the cut-off bias from the pentode so that the circuit continues to oscillate with the same phase as the original shock-excited oscillation.

The coho oscillates at I.F., and comparison with the echo is made at this frequency because it is easier to maintain the required stability than it would be at signal frequency.

The phases of coho and echo signal are compared by means of a phase-sensitive detector, the basic principle of which is shown in Fig. 13.2. In this circuit the connections are such that, whereas the coho signals are fed to the two diodes (via the pentodes) in opposite phase, the echo signals are fed in the same phase.

The two diodes are connected in opposite directions, and the output, in the absence of echo signal, is therefore zero.

When an echo signal is present there will be a resultant output whose magnitude depends only on the phase difference between the coho and echo signals. The signals are separately limited to definite values before application to this circuit, so that their individual initial magnitudes do not affect the result.

The output is fed to a comparison amplifier by two paths, one direct, and the other through a mercury delay tube which introduces a delay exactly equal to the interval between transmitter pulses. The sign of the signal through one of these paths is reversed before the signal is applied to the comparison amplifier, and the final output is therefore proportional to the difference between one echo signal and the next.

Equality between the time delay introduced by the delay line and the interval between transmitter pulses is automatically achieved by using each transmitter pulse, after passing through delay tube, to trigger the next. The mercury tube thus controls the p.r.f. accurately.

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CHAPTER XIV

TIMEBASE CIRCUITS

ACCORDING to the type of display used, timebase circuits are required to produce a suitable waveform of voltage or current for electrostatically or magnetically deflected tubes, respectively. Fundamentally, all timebase circuits depend on the time constant of a circuit containing resistance and capacitance or inductance.

With most displays the voltage or current is required to rise uniformly from zero to a maximum, although in special cases some other law of rise may be required.

The start of the timebase must be synchronized with or by the firing of the transmitter, and the circuit must revert to its initial state very shortly after the completion of the sweep. Control circuits for starting and stopping timebase circuits are dealt with in Chapter V.

Simple Timebase Circuit for A-Type Display

In the circuit of Fig. 14.1, V_1 is normally conducting heavily because its grid is connected to H.T.+ through R_1. C_2 is therefore held discharged.

If a negative switching wave, of sufficient amplitude, is applied through C_1, V_1 is cut off, and C_2 starts to charge through R_2. The voltage across C_2 rises exponentially, but the output can

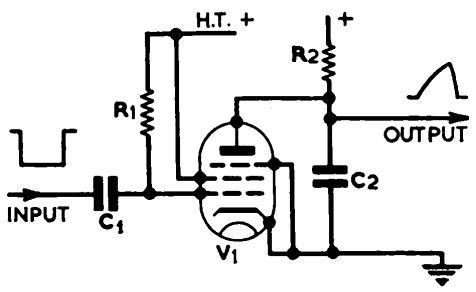

FIG. 14.1.—SIMPLE TIMEBASE FOR "A" TYPE DISPLAY.

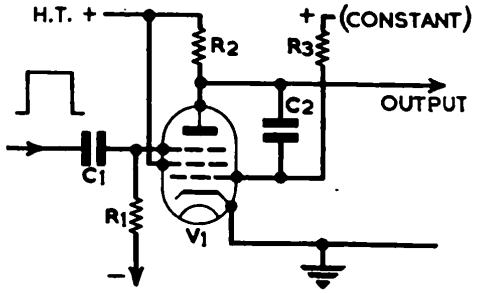

FIG. 14.2.—MILLER TIMEBASE CIRCUIT.

be kept approximately linear by confining the working part of the sweep to a comparatively small fraction of the total charging voltage. This has the disadvantage that, either the output voltage will be low, or a very high voltage must be used for charging. This objection can be overcome by using improved circuits, of which the Miller and the Bootstrap are examples.

Miller Timebase Circuit

This circuit, used extensively for the production of linear timebases, is shown in Fig. 14.2. V_1 is initially cut off by

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negative bias applied to its suppressor grid. The anode is therefore at H.T.+ potential.

If a positive switching wave is applied to the suppressor through C_1, the valve conducts and its anode voltage falls sharply due to drop in R_2. This drop is communicated to the control grid through C_2, which action tends to oppose, but not entirely neutralize, the fall in anode voltage. A state of temporary stability is rapidly reached where the two effects balance one another. The timebase waveform starts at this point.

C_2 starts to charge through R_3, and as the grid voltage rises that of the anode falls in the normal way. If X is the effective stage gain, a rise of one volt on the grid is accompanied by a

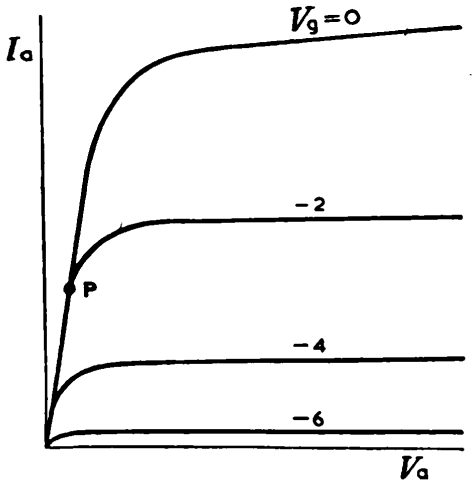

FIG. 14.3.—PENTODE CHARACTERISTICS.

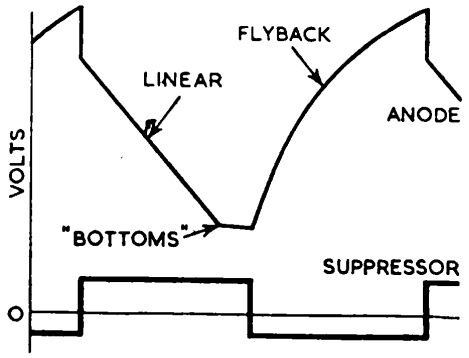

FIG. 14.4.—MILLER TIMEBASE WAVEFORMS.

fall of X volts at the anode; the total voltage across C_2 therefore changes by $X + 1$.

Comparing this circuit with that of Fig. 14.1, it will be seen that, in effect, the capacitance of the capacitor has been increased $X + 1$ times, and the output waveform voltage X times. This is virtually equivalent to working over only a small part of the charging curve, thereby securing substantial linearity, without the attendant disadvantage of low output voltage.

The process continues until the valve " bottoms ", i.e. reaches some point such as P (Fig. 14.3) on the characteristic curve where further rise of grid voltage produces little or no further change in anode voltage. This is the end of the useful part of the sweep.

Removal of the switching wave allows the circuit to return to its initial state in readiness for the next sweep. The waveform is shown in Fig. 14.4.

If sufficient voltage sweep has been obtained before the valve bottoms, the switching wave may, of course, be removed then.

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Bootstrap Circuit

This circuit, of American origin, is shown in Fig. 14.5.

In the quiescent state, the grid of V_1 is held positive through a high resistance R_1 and the valve is bottomed, its anode being nearly at earth potential. C_1 is practically discharged.

When the timebase is required to start, V_1 is cut off by the application of a negative switching wave to its grid. C_1 starts to charge through R_2. The rise of voltage across C_1 is communicated by cathode follower action of V_3 through C_2 to the top end of R_2. This causes diode V_2 to cut off.

If the capacitance C_2 were infinitely large, and if V_3 acted as a perfect cathode follower, the rise of voltage at the top end of R_2 would be exactly equal to the rise at the bottom end, i.e. the charging current of C_1 through R_2 would be constant, and a truly linear timebase waveform would be produced. In practice, these assumptions of perfection do not hold good, but the circuit nevertheless gives very satisfactory results.

Cessation of the negative switching wave allows the circuit to revert to its initial state.

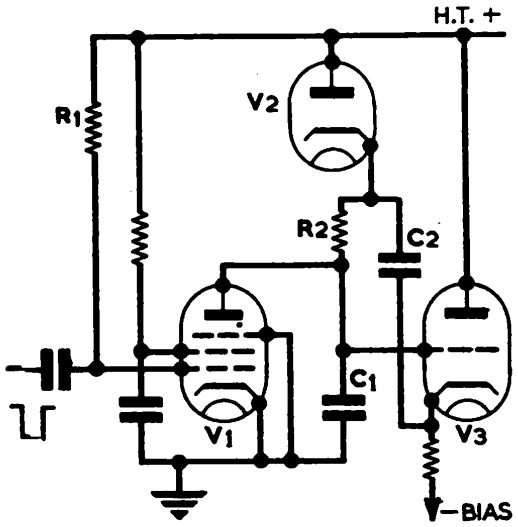

FIG. 14.5.—BOOTSTRAP CIRCUIT.

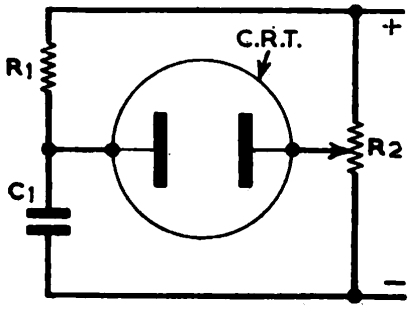

FIG. 14.6.—TIMEBASE WITH BALANCING POTENTIOMETER.

Timebase employing Potentiometer

Fig. 14.6 shows the rudiments of a circuit which can be used when accurate ranging is required.

The timing circuit consists of R_1 and C_1, which can be switched by any convenient method (not shown), such as that of Fig. 14.1. The upper end of C_1 is connected to one of the X-plates of the cathode-ray tube, the other plate being connected to the slider of the potentiometer R_2. The total range of voltage across C_1 is made much greater than is needed to deflect the spot right across the screen, and the slider of R_2 is adjusted to oppose the voltage on C_1 and bring the leading edge of the wanted echo to the centre of the screen, indicated by a cross-wire, as shown

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in Fig. 13.1 (ii). Range is then read off from a scale attached to the slider of the potentiometer R_2, which is accurately calibrated in yards, or other convenient unit. As this is a "null" method, linearity of timebase voltage is not necessary, but the potentiometer must be calibrated to fit the actual law by which the voltage rises.

Timebase having Speeded-up Portion

When it is required to examine a portion of the timebase sweep in greater detail, that part can be speeded up by using the circuit of Fig. 14.7. In the quiescent state V_1 takes all the current because its grid is connected to H.T.+ and R_4 is large. V_2 is therefore cut off.

If the normal linear timebase voltage is applied, in the negative direction, to the grid of V_1, a point is reached, at some time depending on the setting of R_5, when V_2 starts to conduct and shortly afterwards takes all the current, V_1 cutting off.

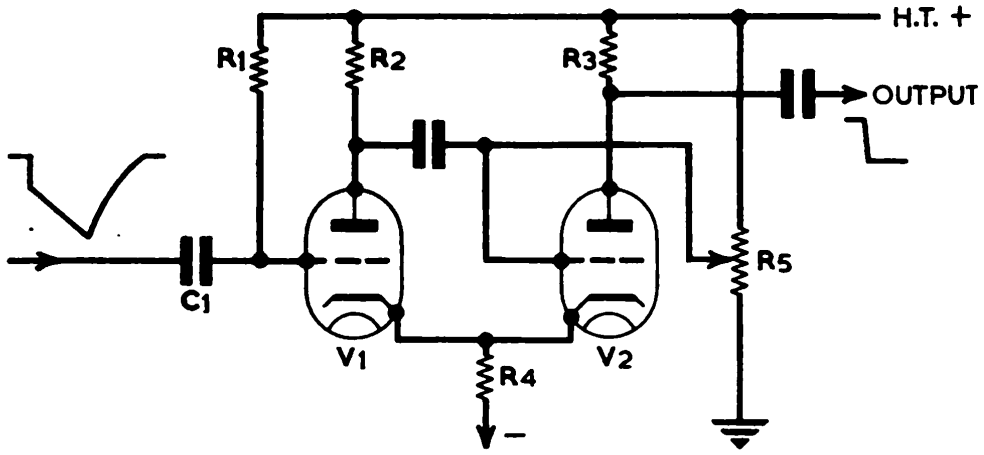

FIG. 14.7.—TIMEBASE WITH SPEEDED-UP PORTION.

This causes a steep negative voltage step at the anode of V_2 which can be used to trigger off a fast timebase circuit. The voltage from the fast timebase circuit can be applied to one deflector plate of the cathode-ray tube, the normal timebase voltage being applied to the other.

The point at which the fast portion occurs is controlled by the setting of R_5, and its duration can be controlled by a diode or other means of limiting the voltage rise on the fast timebase circuit.

It is desirable to feed V_1 through a cathode follower buffer stage to avoid distorting the main timebase waveform due to the loading effect of V_1.

Spiral Timebase for J-type Display

This can be produced by deriving a rotating field, of decreasing amplitude, by phase-splitting the output from a damped

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oscillatory circuit. (See section on Goniometer in Chapter XV for method of deriving rotating field.)

Timebase Circuits for Magnetically Deflected Tubes

Since the deflecting coil has inductance and resistance, it is necessary, in order to produce a linearly rising current, to apply to the coil a voltage waveform consisting of an initial step followed by a linearly rising portion.

Whilst circuits can be devised for the direct generation of such a waveform, it is usual in practice to employ a circuit giving a waveform which is roughly correct and to rely on negative feedback, in a suitable circuit, to constrain the current waveform to the required shape.

Example of Timebase for P.P.I. with Magnetic Tube

In the circuit shown in Fig. 14.8 the main timing circuit consists of R_1 and C_1. In the quiescent state, C_1 is held discharged because V_1 is conducting, its grid being connected to H.T.+ through the high resistance R_3.

When the transmitter fires, a negative square wave is applied

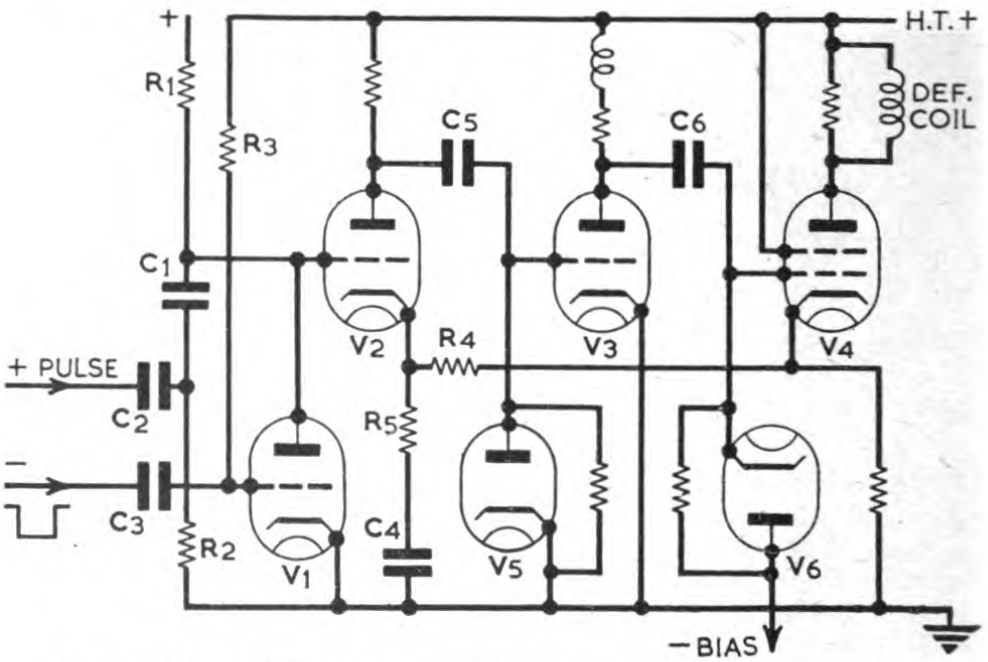

FIG. 14.8.—TIMEBASE FOR P.P.I. USING MAGNETIC C.R.T.

to the grid of V_1, cutting it off. C_1 thereupon starts to charge, and the rising voltage across it, amplified by V_2 and V_3, is applied to the power tetrode V_4, which has the deflecting coil in its anode circuit.

The time constant of R_1C_1 is several times larger than the required sweep time, and the voltage across C_1 therefore rises nearly linearly. Heavy negative feedback via R_4 limits the amplification and ensures that the voltage at the cathode of V_4,

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which is proportional to the current in the deflector coil, is very nearly equal to that applied to the grid of V_2. The current in the coil therefore rises substantially linearly, proportionately to the rise of voltage across C_1.

When the switching wave is cut off by the controlling circuit (usually a multivibrator) reverting to its initial state, V_1 again conducts and discharges C_1. The diodes V_5 and V_6 limit the amplification in the reverse direction, preventing too rapid cutting off of V_4 and limiting the high voltage inductive kick across the coil.

A positive pulse applied through C_2, which is very small, at the start of the sweep provides a small initial kick to compensate for self-capacitance of the coil, etc. This is also assisted by the circuit R_5C_4 (of small time constant) short-circuiting the feedback for a short instant at the start.

It should be noted that, although V_4 is initially cut off, it rapidly reaches conduction-point at the beginning of the sweep because there is no feedback until current starts to flow in V_4.

P.P.I. Timebase—Another Example

In the circuit of Fig. 14.9, when it is in the quiescent state, all the electrodes of V_3 and the grid of V_5 are held at a suitable potential below earth by negative bias. V_5 is cut off. (Note the absence of H.T. supply to all valves except V_5, and the cathode of V_1.)

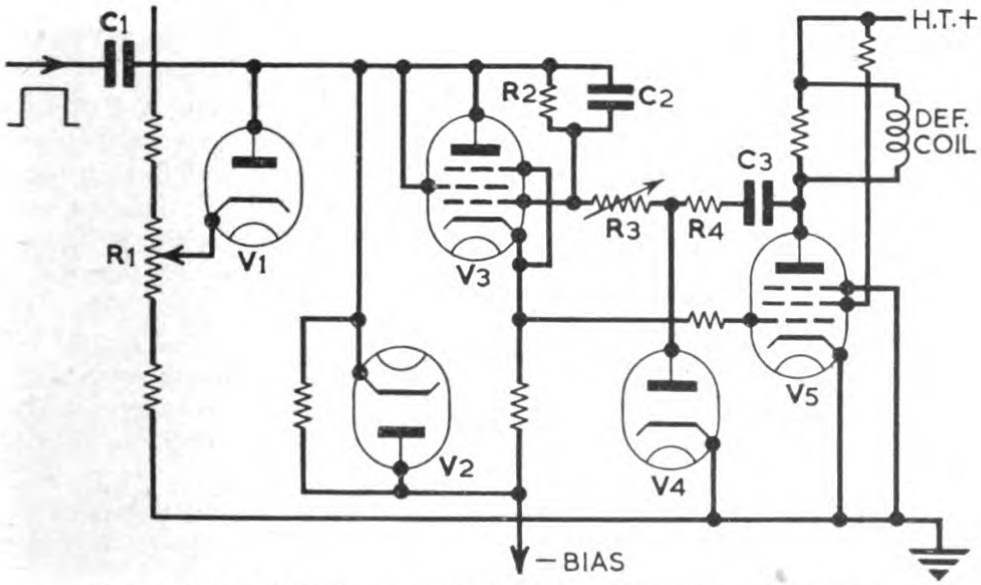

FIG. 14.9.—TIMEBASE FOR P.P.I.—ANOTHER EXAMPLE.

When a positive switching wave is applied through C_1, the anode and grid potentials of V_3 rise to a voltage determined by the setting of R_1. The cathode follows, and with it the grid of V_5, which starts to conduct.

Linearity is substantially achieved by the feedback circuit

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$R_2R_3R_4C_3$, whose time constant is adjusted to approximate equality with that of the deflecting coil.

Assume that the correct waveform of voltage exists at the anode of V_5; the equality of time constants means that the current in $R_2R_3R_4$ will be constant (ignoring the effect of C_2, which is small). From this it follows that the voltage on the grid of V_3 will also be constant. Actually the voltage on the grid of this valve must rise slightly during the sweep, but as the amplification of V_5 is large, it is necessary for the output waveform to deviate from the theoretically correct form by only a very small amount to permit the necessary rise in grid voltage of V_3 to take place. Any *appreciable* departure will alter the current in $R_2R_3R_4$ sufficiently for the change in drop across R_2 to correct the error. The capacitor C_2 across R_2 is small, its function being to pass a small initial pulse to compensate for stray capacitances.

In practice, R_3 is adjusted by trial. The rate of rise, and therefore the range of the timebase sweep, depend on the setting of R_1.

When the control multivibrator reverts to its initial state and the input switching wave is cut off, the output valve V_5 again cuts off. Diode V_3 prevents overshooting, and clamps the potentials ready for the next sweep. Diode V_4 assists in absorbing the inductive surge due to cessation of current in the deflection coil. This diode is non-conducting at other times because the junction of R_3 and R_4 is always below earth potential during the sweep.

Balanced Waveform

If a magslip (which is a transformer) is used in a timebase circuit, with or without the addition of a conventional transformer, in order to avoid having to rotate the deflecting coil of a P.P.I., it is necessary that the waveform be balanced with respect to zero. This is because a transformer cannot pass a d.c. component. It is not necessary that the positive and negative half-cycles be of the same shape, but they must have the same area.

Fig. 14.10 shows a circuit for producing a balanced waveform. In the quiescent state V_1 is conducting and the relative values of R_1 and R_2 and the setting of R_3 are such that V_2 is cut off on its control grid.

The timebase sweep is started by the application of a positive switching wave to V_1 via C_1. This makes V_1 non-conducting and leaves the grid of V_2 free to change its potential. As the grid is connected to H.T.+ through R_1, an immediate rise of voltage occurs. This is followed by the normal Miller timebase sweep.

At the end of the sweep, when V_2 is again cut off, the inductance of the primary of T_1 starts to oscillate with associated capacitances (including strays) and the first half-cycle of oscillation takes the anode of V_2 above H.T.+. During this period

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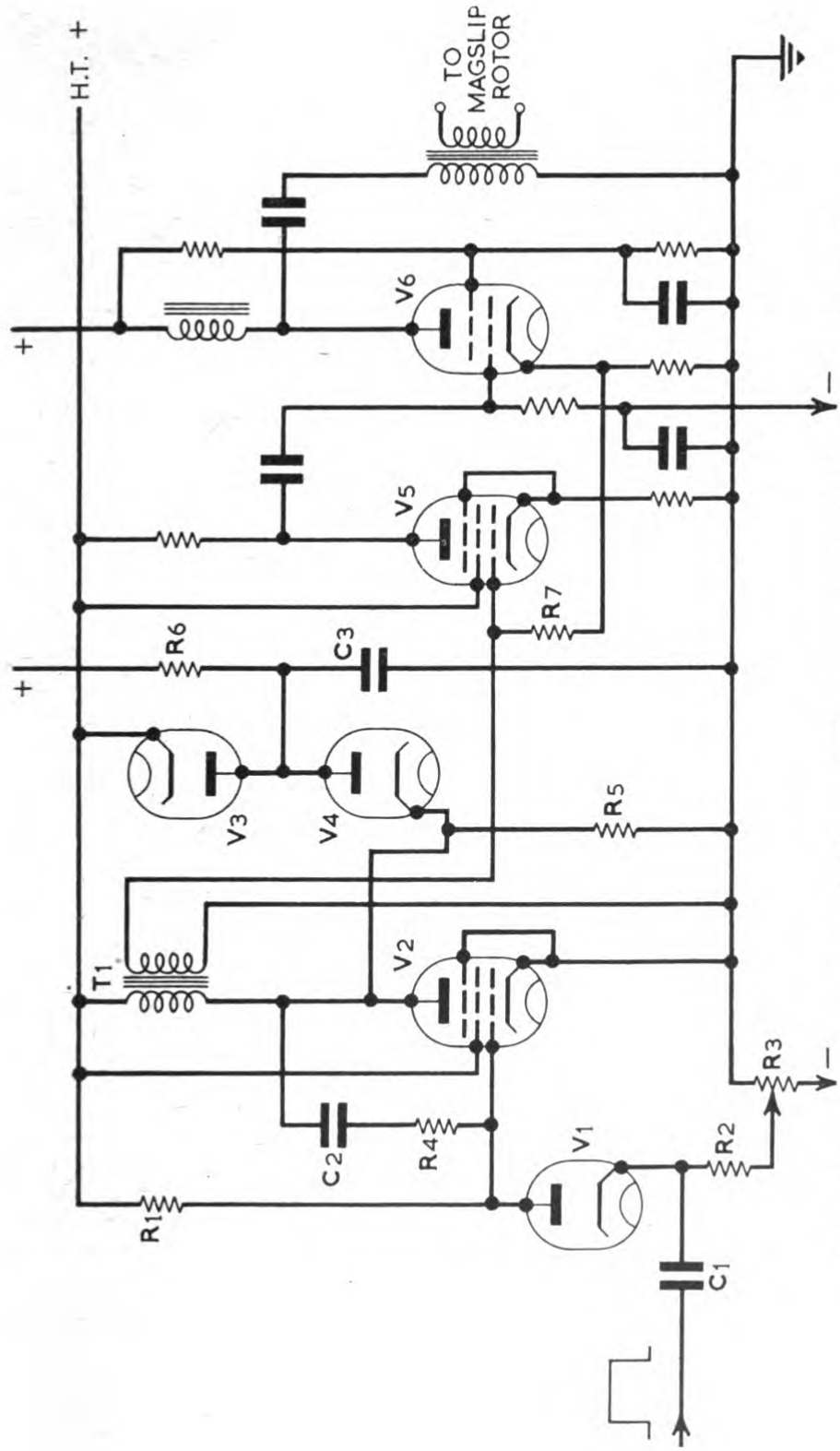

FIG. 14.10.—CIRCUIT FOR PRODUCING TIMEBASE WITH BALANCED WAVEFORM.

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V_4 is non-conducting because, during the Miller rundown, the cathode (and anode) have followed the anode of V_2. By the time the negative half-cycle starts, C_3 has charged up through R_6 (connected to a voltage above that of the main H.T.) sufficiently for V_3 and V_4 to conduct. The oscillation is therefore rapidly damped out, leaving the anode of V_2 at H.T. potential.

As the initial and final voltages across T_1 are both zero (neglecting the negligibly small effect due to resistance of the winding), and as the mean voltage across an inductance in a circuit operated cyclically must be zero, it follows that the waveform of voltage across T_1 is balanced.

The output voltage waveform is fed to the output stages V_5 and V_6, the heavy negative feedback via R_7 forcing the output current waveform to follow the input voltage almost exactly. The output *current* waveform is therefore also balanced, and starts and finishes at zero, as required.

If the cathode of V_3 is connected to a point whose voltage is slightly different from the main H.T. it is possible to obtain some compensation for imperfections due to hysteresis in the core of the deflecting coil, and other things.

A small negative voltage fed to the grid of V_5 at the start of the sweep through a small capacitor (not shown) will assist in minimizing the effects of stray capacitance in coil, magslip windings, etc.

An alternative method of dealing with the necessity for a balanced waveform where transformers are used is to employ a normal sawtooth waveform and accept the fact that, by the time this has passed through a transformer, the zero line will have been automatically shifted to eliminate the d.c. component.

This means that the trace will start, not at the centre of the screen, but at some point at the edge on the side opposite to that on which it is required. The trace must therefore be blacked out until the spot reaches the centre, and the transmitter pulse must be delayed until this instant. It is therefore necessary to trigger the transmitter from the timebase circuit, with appropriate delay, or control both transmitter and timebase by a special synchronizing circuit.

Centre-expand Circuits

A small steady current fed to the deflecting coil from the H.T. supply through a high resistance will provide for an expanded centre display. This method is impracticable where there are transformers in the circuit, and in this case the desired result can be obtained by superimposing a square waveform on the main timebase waveform. The length of the relevant half-cycle must be great enough for it to overlap the active part of the sweep at both ends.

Special Timebases

Timebases following special laws may be required for some purposes. An interesting example is the case of an airborne

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P.P.I., where the difference between slant range (which is what the radar measures) and the horizontal or plan distance along the ground is appreciable at high altitudes. This causes distortion of the P.P.I. picture.

The waveform theoretically required to correct the distortion can be approximated to by superimposing one or more exponential timebases, of suitable time constants, on a linear timebase. The superimposed waveform will be different for different heights above earth, and its start must be delayed by a time equal to that required for a signal to travel to earth vertically and back again. This is necessary in order that the start of the sweep may correspond to zero plan range.

Somewhat complicated circuits have been devised to perform the necessary adjustments for different heights automatically. Exact compensation is not possible, but a very fair approximation can be obtained.

A simpler, but less accurate, method is to use a linear timebase of slightly modified rate of rise, and delay its start by about half the theoretical amount.

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CHAPTER XV

CALIBRATOR AND STROBE CIRCUITS

THE length of the trace on a cathode-ray tube is too small to enable the range of a particular echo to be measured with any great degree of accuracy. For accurate ranging the null method described on page 134 may be used. Alternatively one part of the trace can be speeded up (as described on page 135) and a marker or markers, representing known ranges, added. The range of the wanted echo is then determined by reference to one or more markers.

The markers may be a series of calibrator pips, spaced at known equal intervals as in Fig. 15.1 (*a*), or a single marker or "strobe" may be produced. This may take the form of a step

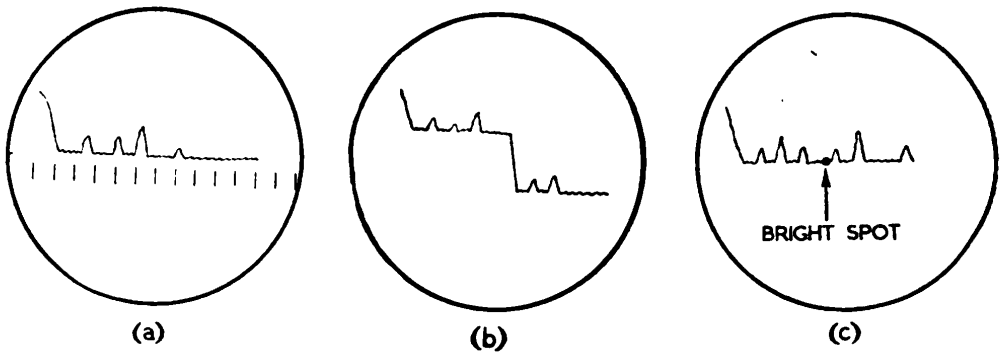

FIG. 15.1.—MARKERS OR STROBES.
(*a*) Calibration pips. (*b*) Step strobe. (*c*) Bright spot strobe.

[Fig. 15.1 (*b*)] or a bright spot [Fig. 15.1 (*c*)]. The former is usually easier to see, but is not applicable to the case of a P.P.I. To measure range, the strobe is moved until it is in line with the leading edge of the relevant echo, and the range is then read off a calibrated scale attached to the potentiometer controlling the position of the strobe along the timebase.

With a P.P.I. the bright spot strobe appears as a circle, concentric with the centre of the screen, and of radius proportional to the range.

A variation of the calibrator pip method consists of means for moving the whole series of pips, *en bloc*, along the trace by a known controllable amount. Any particular pip can then be aligned with a selected echo.

Circuits for Generating a Series of Calibrator Pips

The basic principle in producing calibrator pips is to make use of an oscillator of accurately controlled frequency, and

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square and differentiate its output. The initial oscillation may be obtained from a crystal or from a " ringing circuit ". An example of the latter is shown in Fig. 15.2.

In the quiescent state V_1 and V_2 are conducting, and V_3 is cut off by negative bias on its grid. When the transmitter fires, a negative-going square wave from a multivibrator controlled by the transmitter is applied to the grid of V_1, causing this valve to cut off. When this happens, the ringing circuit L_1C_1 starts to oscillate at its natural frequency. The oscillatory voltage, amplified if necessary, is applied to the grid of V_2 which it overdrives sufficiently to produce a square wave at the anode, the first half-cycle being positive. This is differentiated by C_2R_1 and a positive pip is produced at the grid, and

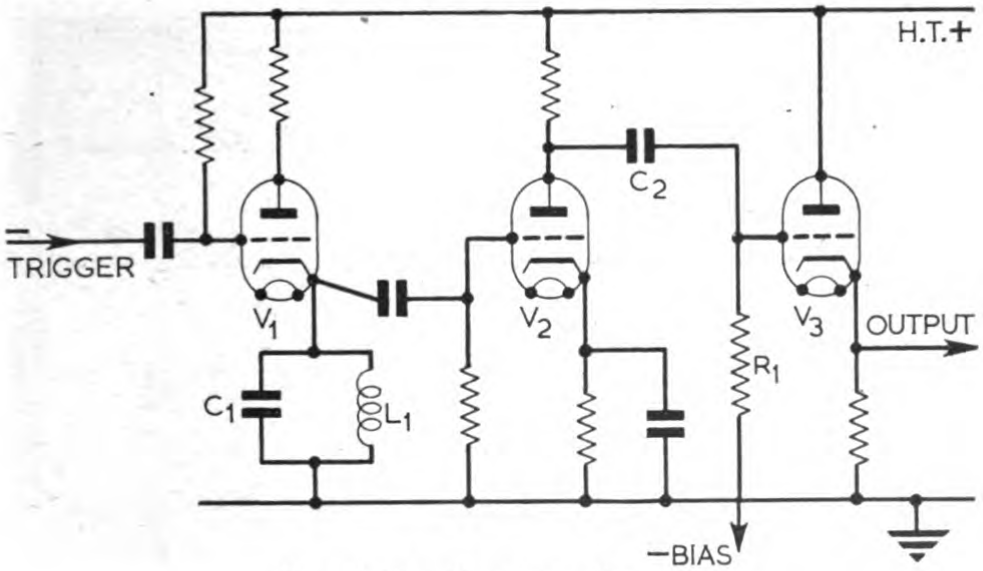

FIG. 15.2.—RINGING CIRCUIT.
For producing calibration pips.

therefore at the cathode also, of V_3 for every cycle of oscillation of the " ringing circuit ".

The first pip is substantially coincident in time with the cutting off of V_1. The negative-going pips, occurring one half-cycle after the positives, have no effect since V_3 is normally cut off.

On the cessation of the negative input wave, the circuit reverts to its initial state.

The accuracy of circuits of this type depends on the maintenance of the natural frequency of the ringing circuit. For more accurate work a crystal may be used, either to check the tuning of the ringing circuit periodically, or to act as the initial source of oscillation.

Circuit of Fig. 15.3

In the quiescent state $V_1V_2V_3$ are conducting and V_4V_5 cut off by negative bias on their grids. At the beginning of the

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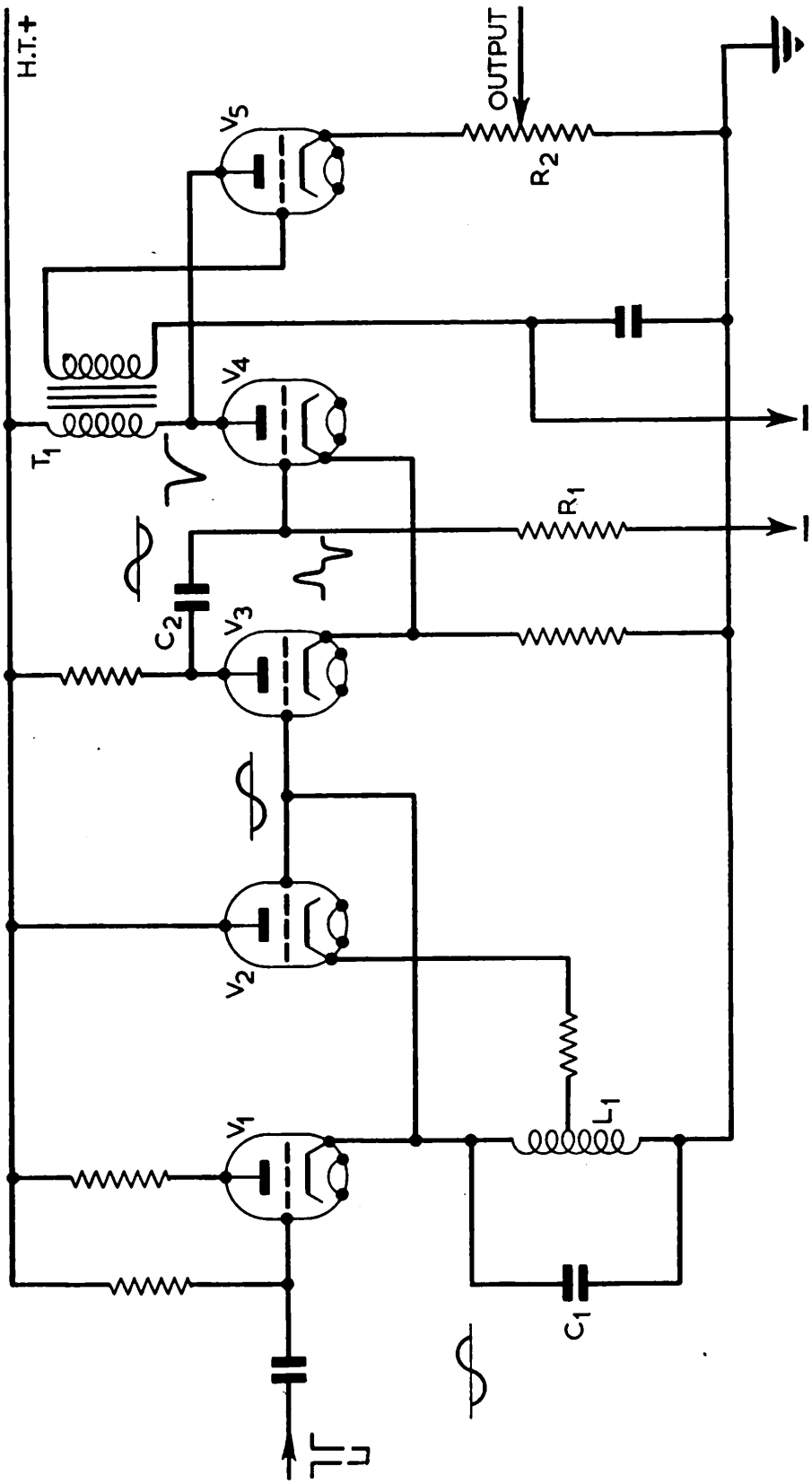

FIG. 15.3.—RINGING CIRCUIT USING BLOCKING OSCILLATOR.

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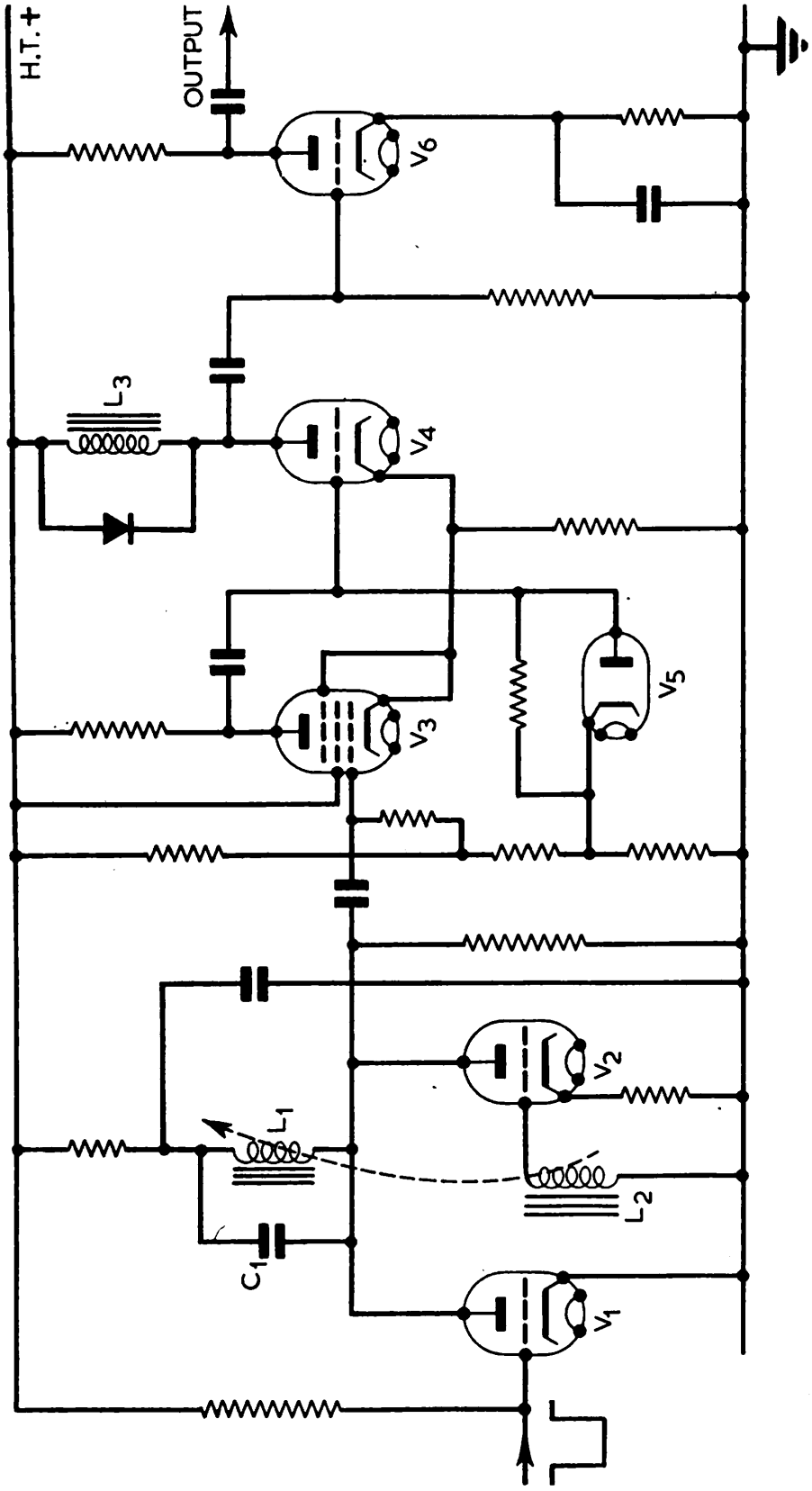

FIG. 15.4.—RINGING CIRCUIT.—ANOTHER EXAMPLE.

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timebase sweep a negative switching wave cuts off V_1 and allows the circuit L_1C_1 to start ringing. The first half-cycle is negative going and cuts off V_3 and allows V_4 to conduct. V_3V_4 form a cathode-coupled flip-flop. T_1 and V_5 constitute a blocking oscillator which produces a sharp pulse of current in V_5. This provides the calibration pip, which is tapped off the cathode resistor R_2. The time constant of C_2 and R_1 is very short, and the flip-flop almost immediately reverts to its initial state, preventing the blocking oscillator from producing another pip until triggered again by the ringing circuit one cycle later. The second half-cycle of the ringing circuit, which is positive going, has no effect because V_3 is already conducting. The function of V_2 is to provide regeneration to prevent too rapid decay of the oscillation in L_1C_1 during the timebase sweep.

Circuit of Fig. 15.4

In the quiescent state $V_1V_2V_3V_6$ are conducting and V_4 is cut off. V_3 and V_4 form a cathode-coupled flip-flop. L_1C_1 form the ringing circuit, regeneration being provided by V_2L_2. The low resistance of V_1 when conducting prevents the ringing circuit from oscillating. A negative-going switching wave cuts off V_1 at the beginning of the timebase sweep, and L_1C_1 ring, the first half-cycle being positive going at the grid of V_3. Very soon after the beginning of the second half-cycle, which is negative, the flip-flop changes to its other state and V_4 begins to conduct. The presence of the crystal rectifier CR_1 across its anode load prevents any appreciable voltage from appearing at the grid of the output valve V_6. At the beginning of the third half-cycle, i.e. one cycle from the start, the flip-flop reverts to its initial state and V_4 cuts off, the interruption of the current in L_3 producing a negative calibration pip at the anode of V_6. Any tendency towards the production of a pip of opposite polarity following the wanted one, due to tendency of L_3 to oscillate with stray capacitance, is suppressed by the crystal rectifier.

This cycle of operations is repeated for every cycle of the ringing circuit. The diode V_5 limits the conduction of V_4 to a suitable value.

Use of Goniometer for moving Calibration Pips along Timebase

A goniometer consists of two pairs of fixed coils at right angles, and a search coil which can be rotated inside them, as shown diagrammatically in Fig. 15.5.

An alternating voltage from the ringing circuit, or other oscillator used for calibrating, is fed to the two sets of coils as shown, a resistance being included in each circuit, and a capacitance C in one. Let L be the inductance of each pair of coils, then if R and C are adjusted so that $L/C = 2R^2$ and $2\omega^2LC = 1$, where $\omega = 2\pi$ times the frequency, and R is the value of one added resistor plus one set of coils, the two sets

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of fixed coils produce a rotating field of constant amplitude. The magnitude of the voltage induced in the search coil is independent of its angular position, but its phase is directly proportional to its angular displacement from some arbitrary zero. The goniometer therefore provides a means of obtaining a calibrator a.c. supply whose phase can be continuously varied. If the output of the search coil is amplified and fed to any of the calibrator circuits described above, the calibration pips can be moved along the timebase by rotating the coil. If a scale is attached, range can be read off by adjusting some particular pip to coincidence with the leading edge of the direct transmitter signal and then moving this pip along until it is aligned with the wanted echo.

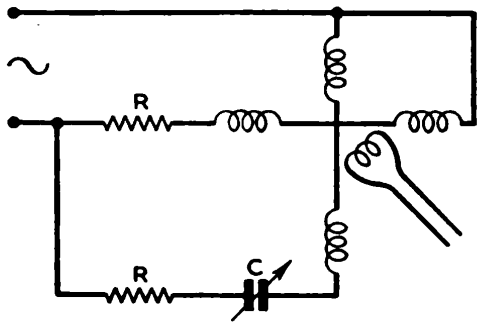

FIG. 15.5.—GONIOMETER.
Used for moving calibration pips along timebase.

Normally the leading edge of the direct signal is not visible on the trace, but this difficulty can be overcome by arranging for the timebase to be triggered a few microseconds before the transmitter fires, or the zero can be set by means of an echo from a fixed object at known range.

If, as will often be the case, a timebase with a speeded-up portion or a strobe, or both, is used, the potentiometers governing the points at which these things occur can be geared to the spindle of the goniometer so that they automatically occur at the correct places.

Crystal-controlled Calibrator

When greater accuracy is required, the ringing circuit of the calibrators described above can be replaced by a crystal-controlled oscillator.

Since the first calibrator pip must coincide with the start of the timebase sweep it is necessary either to pulse the crystal, i.e. start it oscillating, at the beginning of each sweep and stop it at the end, or else arrange for the sweep to start and the transmitter to fire at the same time as a pip.

Although pulsed crystal circuits are possible, their adjustment presents some difficulty, and the second alternative is generally used.

If, as will generally be the case, the p.r.f. of the system is governed by the frequency of the a.c. power supply, the circuit shown in Fig. 15.6 can be used.

V_2 is a thyratron and V_4 a pentode normally cut off on both its control and suppressor grids. During the half-cycle when the upper end of T_1 is positive, C_1 is charged through V_1 and

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R_1. V_2 is held off because its grid is negative due to being connected to the lower end of the transformer, whose winding is earthed at some intermediate point. Calibrator pips, from the crystal circuit, are fed through V_3 to the control grid of V_4, but have no effect because this valve is cut off on its suppressor.

Shortly before the end of the half-cycle, when the grid of the thyratron is approaching zero voltage and going positive, this valve fires, and the current through R_2 raises the potential of the suppressor of V_4, allowing this valve to conduct when pips are applied to its control grid.

Negative-going calibrator pips are therefore produced at the

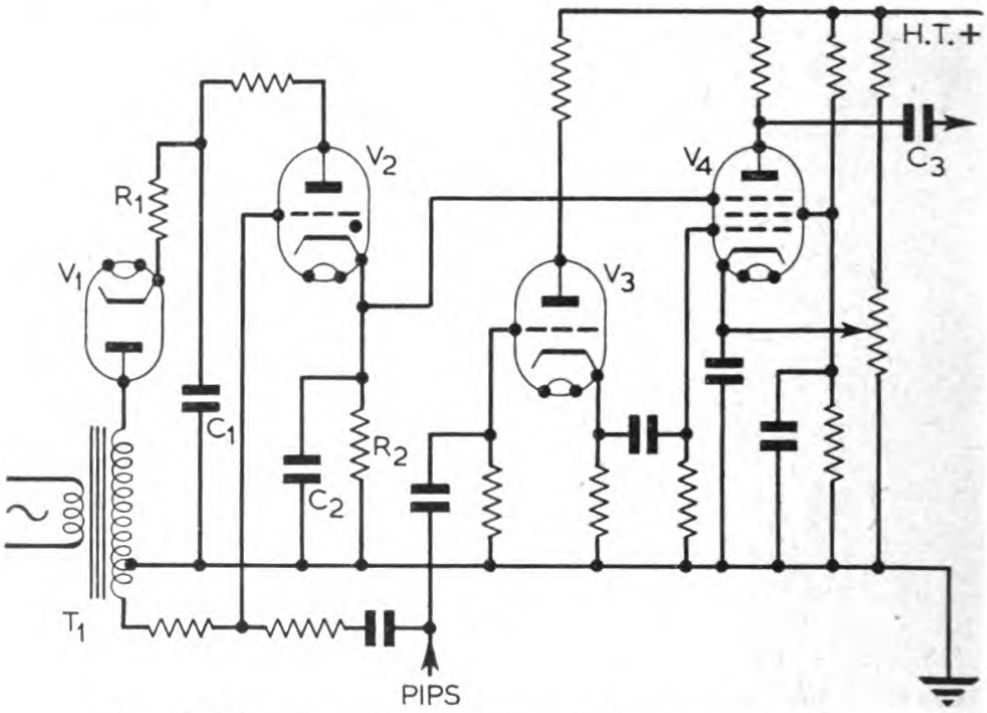

FIG. 15.6.—CRYSTAL-CONTROLLED TRIGGER CIRCUIT.

output (C_3). The first of these pips is used to trigger the modulator and timebase circuits, ensuring the required synchronism. C_1 discharges fairly rapidly so that only a few pips reach the output for each cycle of operation.

To avoid possible jitter which might occur if V_2 fired at nearly but not quite the same instant as a pip occurred, it is advisable to superimpose pips on the grid of V_2. This ensures that when the voltage on the grid has nearly reached the point at which the valve would fire in any case, it is then caused to fire by the pip. The time constant of C_2R_2 is made such that the voltage on the suppressor of V_4 does not rise quickly enough for the pip which triggered V_2 to get through V_4. The next pip is the one which triggers the modulator and timebase. To avoid interaction between the two circuits, the cathode follower V_3 acts as a buffer stage.

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Crystal Oscillator Circuits

Circuits commonly used are shown in Fig. 15.7. In order that regeneration may occur and maintain oscillations it is necessary

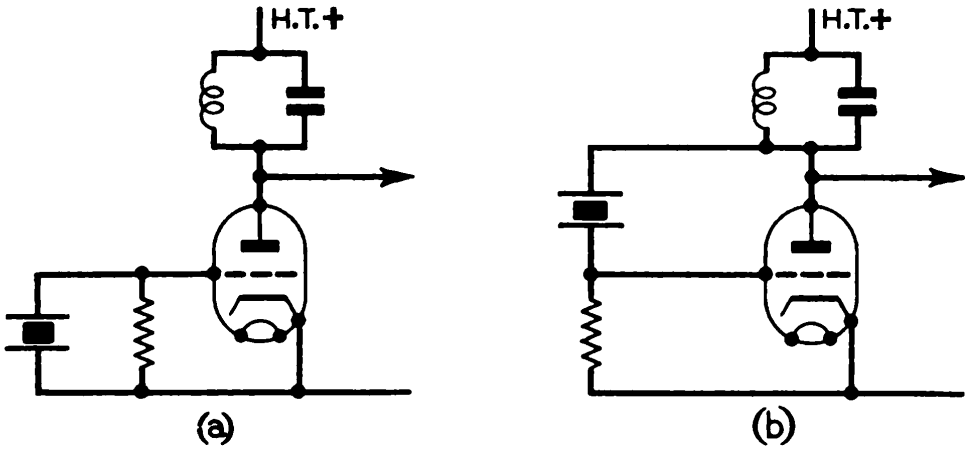

FIG. 15.7.—CRYSTAL-CONTROLLED OSCILLATOR.
(*a*) Grid-cathode connection.
(*b*) Anode-grid connection.

that the anode circuit be tuned to a frequency slightly higher than that of the crystal in case (*a*) and slightly lower in (*b*).

The output contains harmonics, and if a pure sine wave is required, as for example when used with a goniometer phase-shifting circuit, the output must be passed through one or more amplifying stages tuned exactly to the crystal frequency. The circuit shown in Fig. 15.8 may be used, where L_1C_1 and L_2C_2 are tuned slightly above and equal to the crystal frequency respectively. In this case, the screen of the pentode acts as anode for the oscillator part of the circuit.

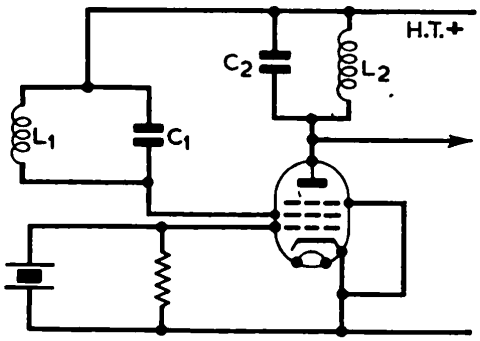

FIG. 15.8.—CRYSTAL-CONTROLLED OSCILLATOR WITH TUNED CIRCUIT FOR ELIMINATING HARMONICS.

Strobe Generating Circuits

Fig. 14.7 is really a form of strobe generating circuit. More complete circuits are shown in Figs. 15.9 and 15.10.

In Fig. 15.9 V_1 and V_2 are initially cut off. Application of a positive square wave to the grid of V_1 starts a Miller time-base, consisting of $R_1R_2C_1V_2$. The anode voltage of V_2 runs down until it reaches a potential determined by the setting of R_3. At this point V_4 begins to conduct and the voltage drop in the associated resistances lowers the potential of the grid

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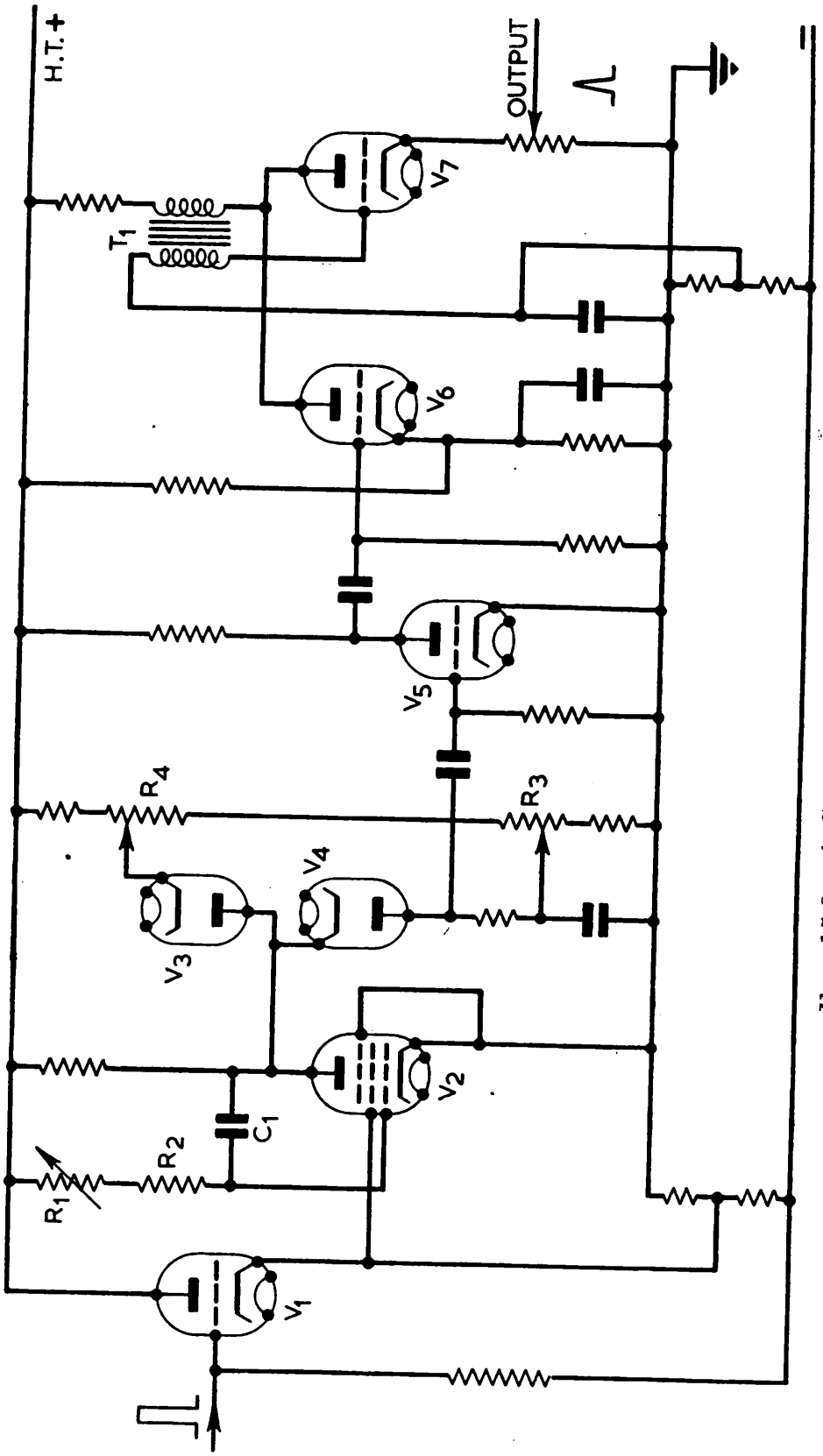

FIG. 15.9.—A STROBE GENERATING CIRCUIT.

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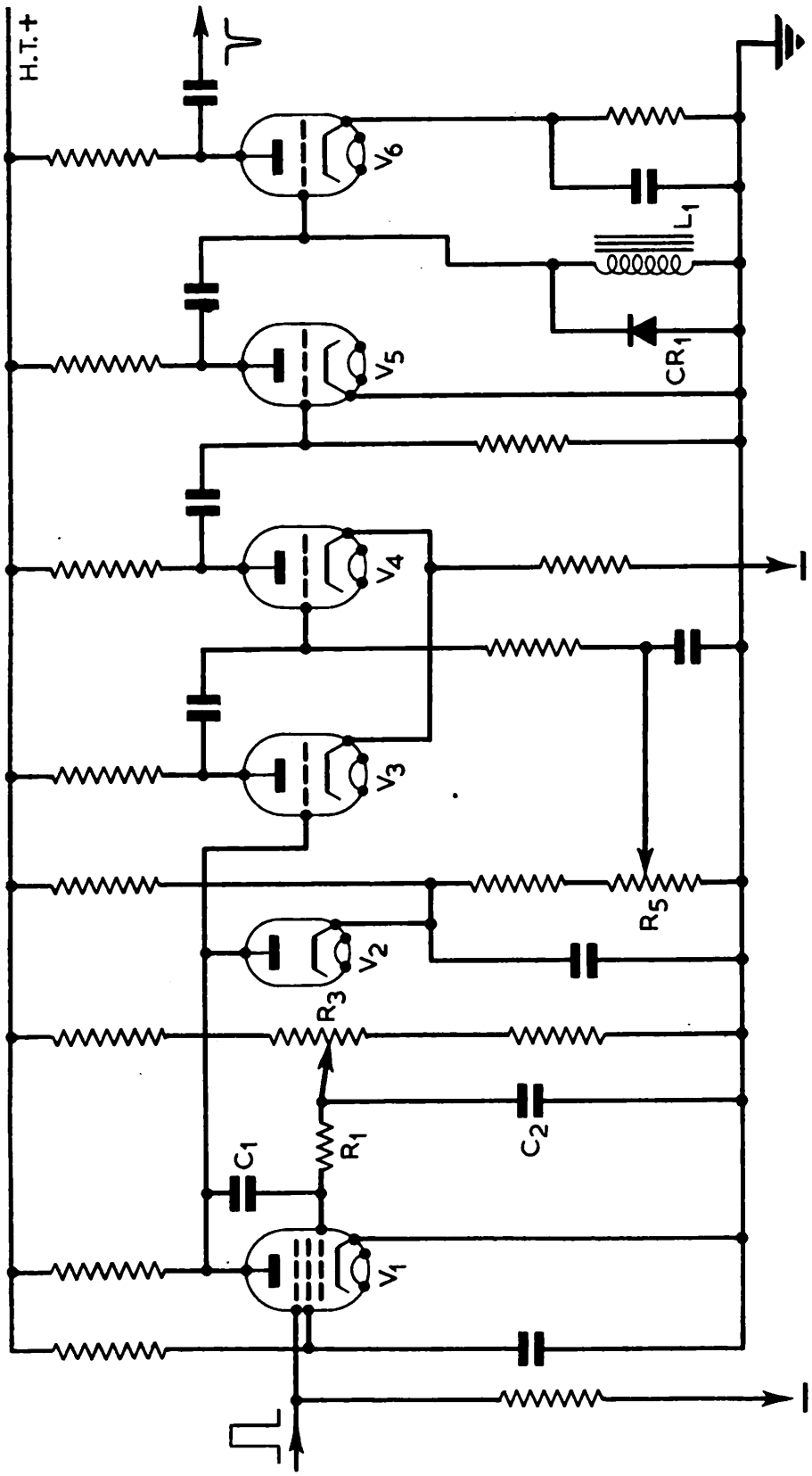

FIG. 15.10.—STROBE GENERATING CIRCUIT—ANOTHER EXAMPLE.

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of V_5, producing a rise in voltage at its anode. This triggers the blocking oscillator T_1V_7 and produces a sharp positive pip in the cathode circuit of V_7. This pip is fed to the cathode-ray tube, where it produces a bright spot on the trace.

If a step strobe is required, the output of V_5 can be arranged to trigger a multivibrator, giving a square wave which can be applied to the deflection plate circuit. R_3 is calibrated to indicate range, and R_4 is a zero adjustment. The duration of the sweep is proportional to the difference of voltage between the two sliders (R_3 and R_4).

In Fig. 15.10 V_1 is initially cut off. V_3 and V_4 form a cathode-coupled flip-flop, V_4 being initially cut off.

A Miller timebase (C_1R_1, etc.) is started, as in the previous case, by a positive switching wave at the input. The rate of fall of potential at V_1 anode is controlled by R_3. As the voltage falls, the cathode of V_3 follows, taking V_4 with it until the latter begins to conduct. This happens at a point determined by the setting of R_5, and triggers the flip-flop, giving positive and negative square waves at the anodes of V_3 and V_4 respectively. Either of these may be used to give a step strobe. In the circuit shown, the negative wave cuts off V_5, producing a positive voltage on L_1 which begins to oscillate at high frequency with stray capacitance. Only one half-cycle occurs, for the next is damped out by the crystal CR_1. A negative-going pip is therefore produced at the output.

If the same circuit is used for speeding up part of the trace and also for producing the strobe, the latter event must be delayed a short time behind the former, in order that it may appear wholly in the speeded-up portion. This can be done by inserting a suitable delay circuit in the feed to the part of the circuit that actually produces the strobe.

The Phantastron

The circuit shown in Fig. 15.11, known as the " phantastron ", can be used for producing a rectangular pulse of known short duration. The trailing edge of this pulse can then be used to trigger another circuit, at a known time interval after the original trigger.

In the quiescent state the anode and grid potentials of the pentode V_1 are clamped to definite values by the diodes V_2 and V_3, screen current is flowing, and the resultant drop across R_5 is sufficient to raise the potential of the cathode to some value higher than that of the suppressor, so that the anode of V_1 is cut off.

The application of a positive trigger pulse to the suppressor allows anode current to flow. This produces a sudden drop of anode voltage, due to R_2, accompanied by a corresponding fall of grid voltage communicated through C_2. The cathode current falls to a value small enough to bring the cathode potential below that of the suppressor, and current continues

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to flow, even although the trigger pulse may have ceased by this time.

V_1 with C_2 and R_3 constitute a Miller timebase, and the Miller

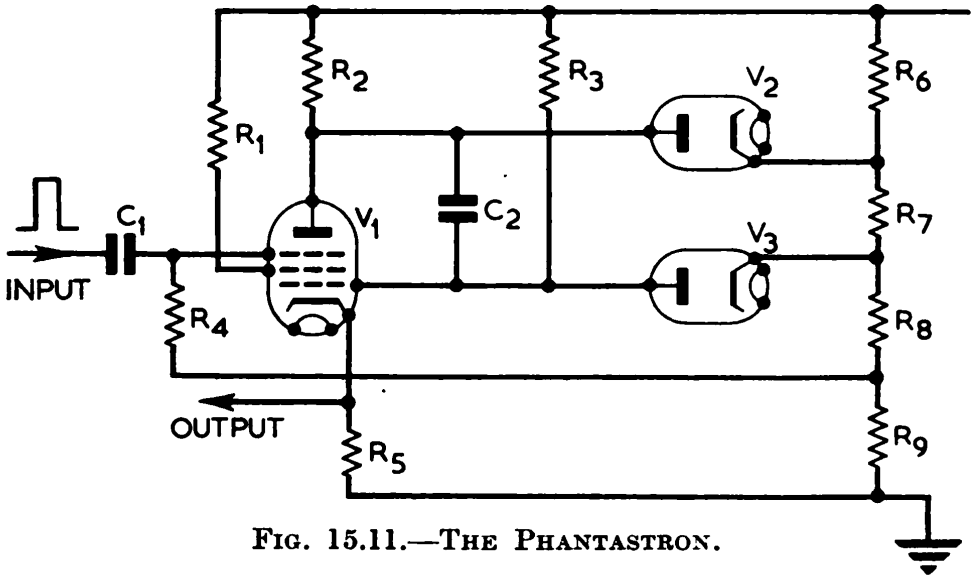

FIG. 15.11.—THE PHANTASTRON.

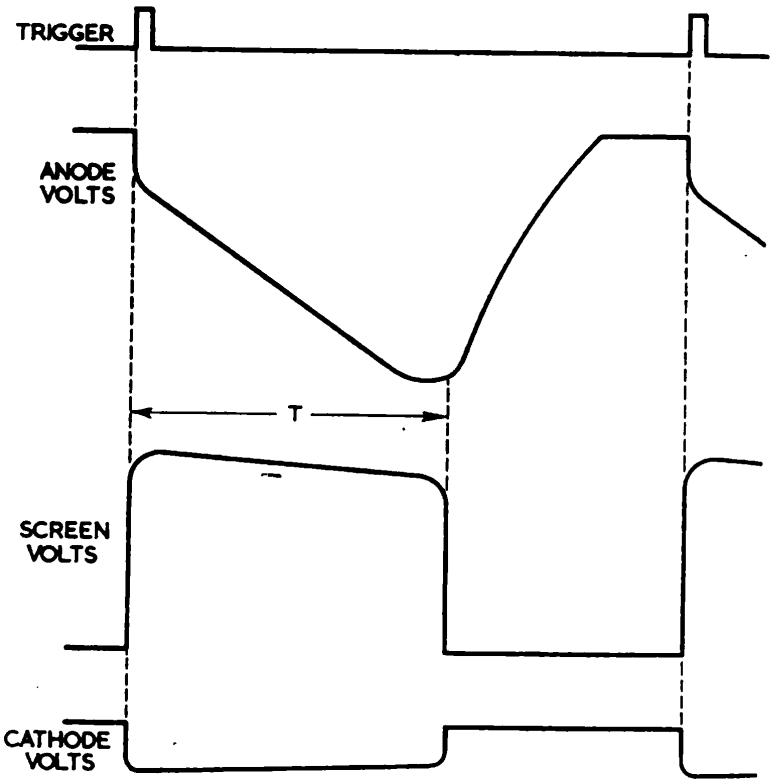

FIG. 15.12.—PHANTASTRON WAVEFORMS.

rundown now begins, the grid voltage, and consequently the cathode voltage, also rising linearly. When V_1 bottoms, the rate of rise becomes greater and a point is rapidly reached at

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which the suppressor again begins to cut off the anode current. The resulting rise of anode voltage is communicated to the grid through C_2, and the action is therefore regenerative, causing the rise in anode voltage to be very rapid until the anode is again cut off. Thereafter the anode returns to its initial voltage exponentially by the charging of C_2 through R_2.

Waveforms are shown in Fig. 15.12. A trigger pulse can be taken from either screen or cathode of V_1, and will be produced at a time T after the application of the initial trigger.

Artificial Lines as Delay Circuits and Calibrators

The artificial line shown in Fig. 15.13 will give a delay between input and output equal to $n\sqrt{LC}$, where n is the number of sections.

If such a line is connected as shown in Fig. 15.14, the " output "

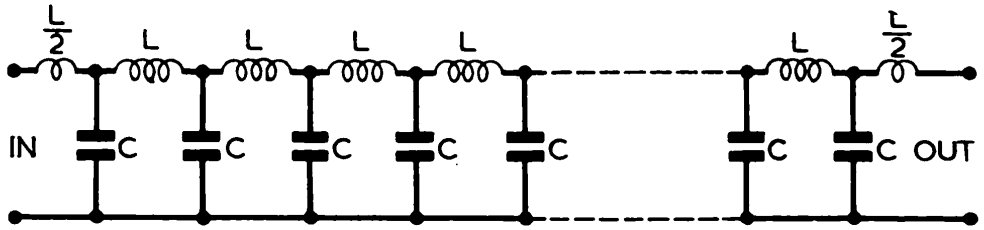

FIG. 15.13.—ARTIFICIAL LINE.

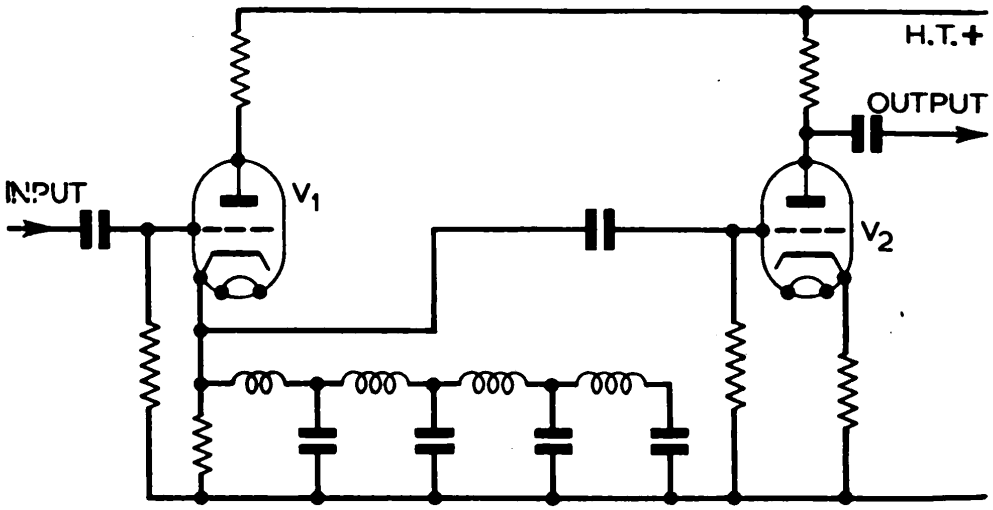

FIG. 15.14.—ARTIFICIAL LINE ARRANGED FOR PRODUCTION OF CALIBRATION PIPS.

end being open, and if there is sufficient mismatch between the line and the cathode resistor, a short sharp pulse applied to the grid of V_1 will be reflected backwards and forwards along the line, producing another pulse each time it reaches the input end, until the energy is dissipated. A series of pulses, spaced $2n\sqrt{LC}$ apart, is therefore obtained at the output of V_2.

This arrangement is an alternative to the ringing circuit.

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Gateing Circuits

If a circuit is required to pass a signal only at certain times in a cycle of operations, a gate can be provided by passing the signal through a pentode which is cut off on its suppressor by suitable negative bias, except when a positive rectangular switching wave is applied to cancel the bias and allow the valve to conduct. Such circuits are required in connection with beam switching or split, with automatic following, the application of A.G.C. to a selected echo, and many other purposes. The required square wave may be generated by some form of multivibrator, or the phantastron may be used.

If, as with automatic following, for example, two gateing waves are required, the second beginning when the first ends, the trailing edge of the first can be used to trigger the second.

Supersonic Delay Cell for Ranging

A vertical tube containing water or other suitable liquid and provided with a piezo-electric crystal at the bottom and a movable plunger at the top can be used for measuring range.

A fraction of the transmitter pulse is fed to the crystal where it is converted into a supersonic pulse. This travels up the tube, is reflected from the plunger, and returns to the crystal at the bottom where it is converted into an electrical impulse.

The time taken for this to happen depends on the velocity of sound in the liquid, and the distance the sound wave has to travel, which depends on the position of the plunger.

The plunger can be moved until the delayed transmitter pulse is coincident in time with the relevant echo signal. From a knowledge of the path length in the tube and the velocity of sound, the range of the target can be calculated. As the velocity of radio waves is roughly 200,000 times the velocity of sound in water, a tube one metre long will suffice for targets up to ranges of about 200 kilometres.

A cathode-ray tube can be used for displaying the two pulses to enable the operator to adjust for coincidence.

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CHAPTER XVI

POWER SUPPLIES

THE main power supply for a fixed radar installation will usually be taken from public mains, and will normally be a.c. at 50 c/s. On board ship or on aircraft, the supply will frequently be d.c., and a motor alternator must be used to provide an a.c. supply suitable for feeding the various power packs in the installation.

If a motor alternator has to be provided, it is very common practice to generate at a frequency suitable for controlling the pulse-repetition rate, say from 500 to 2,000 c/s. In fact, this is frequently done even when the primary supply is a.c.

H.T. Supplies

Power packs are generally conventional types as used in radio, and may have either valve or metal rectifiers.

Where a controlled steady voltage is required a stabilizing circuit must be used. Details of such circuits vary considerably, according to exact requirements in any particular case. The simplest method is to connect a neon stabilizing tube or tubes across the H.T. lines, but if closer control is required a more complicated circuit must be used. Fig. 16.1 is an example. In this circuit V_1 is a power valve which carries the whole of the H.T. current required (two or more valves may be connected in parallel if necessary). The drop across V_1 depends mainly on the voltage on its grid, which is connected to the anode of the pentode V_2. The cathode of V_2 is maintained at a substantially constant voltage by the neon stabilizer V_3, and its grid voltage, being tapped off R_2, is proportional to the output voltage.

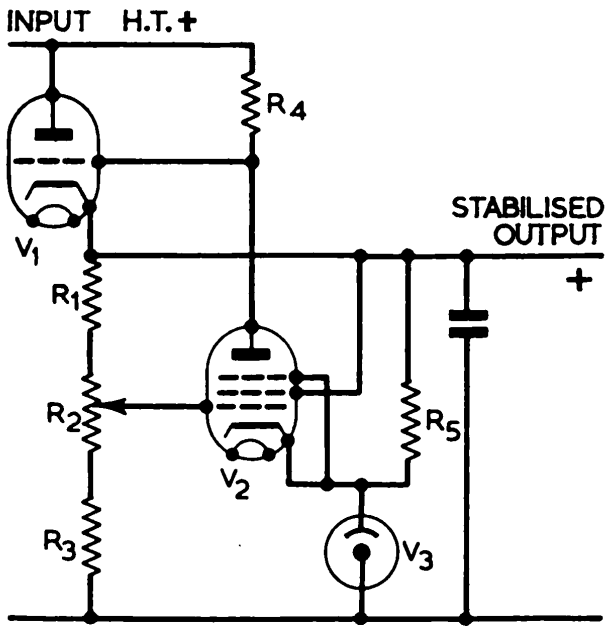

FIG. 16.1.—A VOLTAGE STABILISING CIRCUIT.

Suppose that, for some reason, the output voltage rises. The

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grid will rise by some fraction of the change (about one-half). The anode current in V_2 will increase, and since R_4 is a high resistance (say 20,000 to 100,000 ohms) the voltage on the grid of V_1 will fall by a much greater amount, with consequent considerable increase of voltage drop in V_1. This counteracts the assumed initial rise in output voltage.

The control is " tight " because a small change in the voltage at the grid of V_2 produces a large change in the drop across V_1.

Adjustment of the output voltage to the required value is by moving the slider of R_2.

There are several variations of this type of circuit.

Protective Circuits

Certain protective circuits are desirable to prevent serious damage in event of a minor fault occurring. For example, a spark-gap connected across the magnetron will prevent the voltage from rising to a dangerously high value in event of the magnetron failing to fire. A thermal delay switch in series with the safety-gap may be used to operate a relay and switch off the power if the failure to fire is persistent.

Time-lag devices are used to prevent the application of H.T. to the magnetron and other valves before the heaters have warmed up properly.

In some parts of the circuit, H.T. and bias voltages may be provided by different power packs, in which case it is desirable to provide means for ensuring that the H.T. cannot be applied unless the bias voltage is correct. This can be achieved by suitable relay circuits.

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CHAPTER XVII

TESTING AND TEST GEAR

APART from specific tests carried out for initial adjustment or dealing with a definite fault, regular periodic testing of a radar installation is of considerable importance because experience has shown that overall performance may deteriorate by as much as 10 decibels or more without the operator being aware of the fact.

The tests which can be carried out in the field are necessarily somewhat limited, but they can nevertheless yield useful information on the general condition of the equipment.

Full use should be made of facilities provided for metering currents and voltages at various points in the circuit, and waveforms should also be checked with a monitor oscilloscope, if this is provided.

It is only practicable to give a brief outline of the subject of testing here, and the operator should always carefully observe the working instructions issued by the maker of the equipment, *paying especial attention to safety precautions.*

Overall Performance Test with Echo Box

An echo box coupled to the aerial feeder will absorb energy during the transmitter pulse, and the resulting oscillation built up in the echo box will persist, with decreasing amplitude, after the pulse has ceased. The time taken for the oscillation to decay to negligible amplitude (say noise level), as shown on the display, is a measure of the overall efficiency of the installation.

Coupling to the feeder may be by a small hole, and the echo box may be put out of action when not in use by detuning or by inserting a resistive attenuator in the length of guide connecting the box to the main waveguide.

An alternative is to set up the echo box, with its own small parabolic mirror and dipole, at a distance of a few feet from the aerial. The dipole picks up some of the emitted energy and re-radiates it. This arrangement is suitable for a P.P.I. installation.

For accurate work the echo box should be tuned each time it is used because, owing to the high Q, a slight deviation from correct frequency will make a large difference in the final result. For greater simplicity, where the highest accuracy is not required, the expedient of rotating or vibrating a small paddle or disc inside the cavity is resorted to. This sweeps the tuning over a small range containing all the frequencies which may be encountered, and yields a fair average result.

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Field Strength Measuring Equipment

A dipole set up several wavelengths from the aerial, and connected to a thermocouple and meter, can be used to measure field strength, and so give an indication of the performance of the transmitter and aerial system. This method is very accurate if properly carried out, but practical difficulties of ensuring that the dipole is always set up in exactly the same position, and that echoes from nearby objects do not introduce variable errors, will frequently limit its usefulness.

Wavemeters

Wavemeters for metre waves follow the general lines of those used in communications work. For centimetre waves, resonant cavities (echo boxes) are generally used, in association with a crystal and meter for indicating resonance. Coupling may be by a small hole to a waveguide in cases where the wavemeter is fixed, but more often a coaxial lead having a probe or pick-up

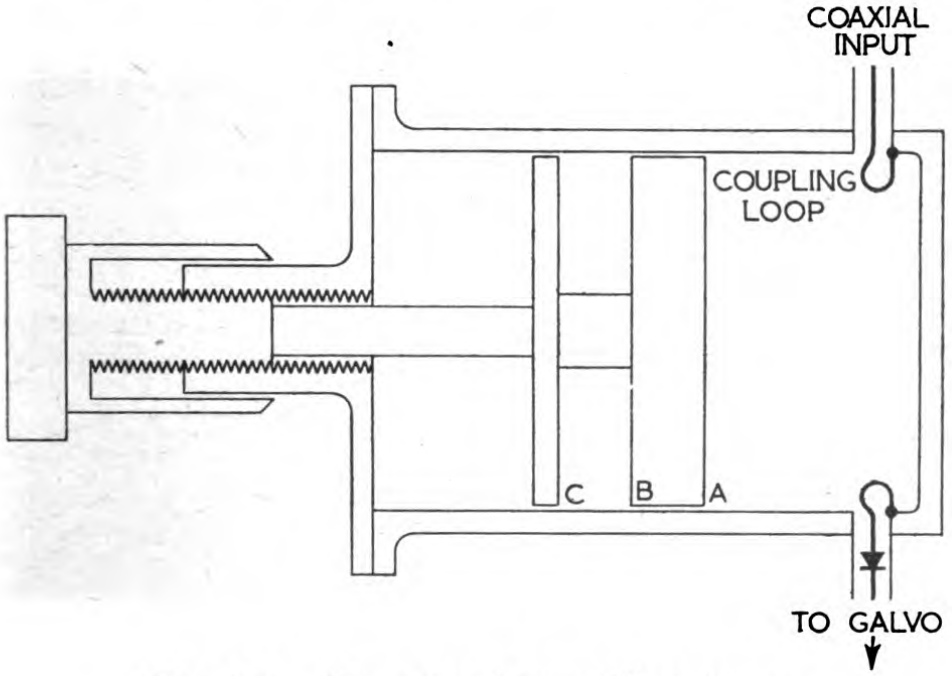

FIG. 17.1.—RESONANT CAVITY WAVEMETER.

loop at its end will be used for coupling to a convenient part of the feeder system.

The general arrangement is shown in Fig. 17.1. The movable tuning plunger carries a micrometer screw from which the wavelength may be read off directly, or by aid of a calibration chart. As sliding contacts are apt to be unreliable, the effect of a direct connection at A without actual mechanical contact between plunger and cavity wall is obtained by using the quarter-wave line principle. The two portions AB and BC each behave as quarter-wave lengths of concentric line, the impedance

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of AB being Z_1, say, and that of BC, Z_2, say. Suppose the impedance between the disc and cavity wall at C is equal to Z. This will be equivalent to Z_2^2/Z at B and $\frac{Z_1^2}{Z_2^2}Z$ at A. Owing to the difference in diameters of the two sections AB and BC, Z_2 is much greater than Z_1 and the impedance at A is therefore very small compared with Z (at C), which is itself reasonably small. Consequently there is practically a short-circuit at A, without actual contact. Strictly speaking, this argument only applies at one frequency, but in practice it holds good sufficiently well over the waveband for which a wavemeter of this type is used, say 10 to 15 per cent each side of mid-point. With careful construction and calibration accuracies better than about 0·1 per cent can be obtained with this type of wavemeter.

Signal Generators and Test Oscillators

A signal generator consists of an oscillator of adjustable frequency, means for setting the output power to a predetermined level, and a calibrated attenuator for feeding an accurately known amount of power to the receiver under test.

As this power may be as low as 10^{-15} watt, the power monitor or setting indicator must be on the input side of the attenuator, and the latter must be accurately calibrated so that the actual power fed to the receiver can be calculated.

For centimetric waves, a piston attenuator (shown in Fig. 17.2) is generally used. The attenuator comprises a length of

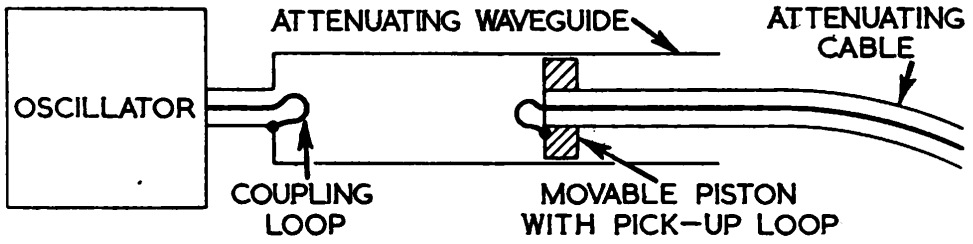

FIG. 17.2.—PISTON ATTENUATOR.
As used with signal generator.

circular waveguide operating below its cut-off frequency and provided with a movable piston carrying a coupling loop.

At frequencies below cut-off, a waveguide introduces a large attenuation per unit length, the amount of which can be calculated from dimensions. The attenuation introduced, being proportional to the length of guide in circuit, is adjusted by moving the piston. By this method, attenuations of 100 decibels can be obtained in a few inches.

The actual frequency, if required to be known accurately, must be determined by a wavemeter, built in or separate.

It is desirable to connect the output of the signal generator to the receiver through a short length of attenuating cable (coaxial cable having its inner conductor made of resistance

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wire), giving a known attenuation, say about 10 decibels. This virtually isolates the receiver from the signal generator in the sense that the output is not affected by the impedance of the load, and the receiver " sees " the impedance of the cable, which can be of suitable value to simulate the aerial.

Square wave modulation, with unity mark/space ratio, is generally used since this simulates the pulses with which the receiver has to deal in practice. It is not practicable to pulse the signal generator on a duty cycle anything like that used for the transmitter pulses, because the *average* power in these circumstances would be far too low, and noise would entirely vitiate the results obtained.

For tests at I.F., signal generators of the type used for communications work can generally be used, because radar I.F.s usually lie between about 20 and 100 Mc/s.

For routine work, where great accuracy is not required, many of the refinements fitted on a signal generator can be omitted. The device is then more commonly known as a test oscillator.

Delayed Pulse Oscillator (D.P.O.)

This consists of an oscillator, with its own small aerial, which emits a pulse at some finite time after receiving a pulse from the transmitter. If set up near a radar set it behaves, in effect, as an artificial echo.

One particular use is testing the recovery time of T.R. cells. The increase in response as the time delay is increased from a very small value gives an indication of the speed of recovery of the cell. With this device the strength of the signal emitted is, of course, independent of the time delay.

Noise Generator

A circuit as shown in Fig. 17.3, where the diode emission is limited by cathode temperature and not by space charge effects, will give, due to random emission of electrons, a noise output whose mean square voltage, between frequency limits f_1 and f_2 c/s, is $3{\cdot}2 \times 10^{-19} R^2(f_2 - f_1)I$, where I is the mean anode current (d.c.). Above about 500 Mc/s the effect of electron

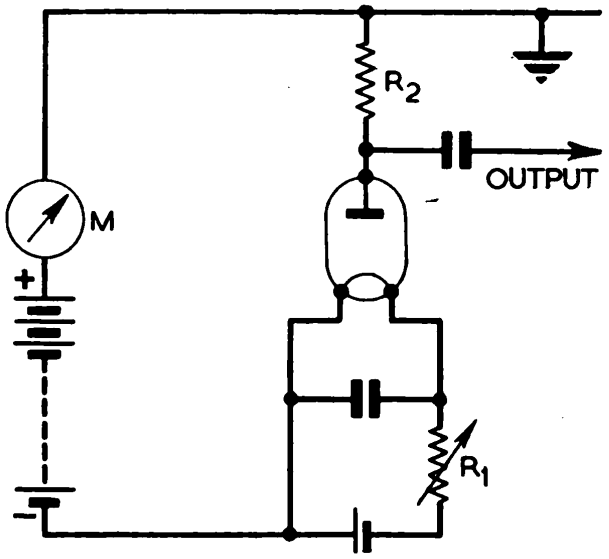

FIG. 17.3.—NOISE GENERATING CIRCUIT EMPLOYING TEMPERATURE LIMITED DIODE.

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transit times causes the above formula to become inaccurate; calibration is therefore necessary if the device is used at higher frequencies.

The noise generator can be used for testing receivers.

Receiver Testing

The overall sensitivity of a receiver can be measured by injecting a known signal from a signal generator or noise source, and measuring the output power on a thermocouple meter or other power-measuring device.

To simulate working conditions it is desirable to leave the aerial connected, and feed in the test signal by loose coupling, such as a probe in the feeder. With a noise source, however, the input power available is usually not sufficient to permit this, and the output resistance of the source must therefore be connected in place of the aerial. This resistance should be made equal to the radiation resistance of the aerial, but if the latter is not known, it is best to match to the feeder, the impedance of which will usually be known.

The real limiting factor in the sensitivity of a receiver is the equivalent noise power. This can be defined as the power of a signal which, when applied to the input, gives an output power equal to that produced by noise alone. The measurement can be made by connecting a signal generator to the receiver, adjusting the gain until the output due to noise (the signal generator being switched off) is some convenient value which can be measured on a thermocouple, and then switching on the signal generator and adjusting the attenuator until the output power is doubled. Provided the signal generator output is matched to the feeder or receiver input, the power delivered by it is then equal to the noise power.

I.F. Alignment and Bandwidth

This operation must be carried out with the A.F.C. disconnected, the output being measured for various frequencies over the required band. Since a series of readings must be taken after making each adjustment, the operation is laborious unless a special signal generator is used. The wobbulator is an example of a suitable generator. In this instrument the frequency is continuously swept over the required band, either by motor drive or in some other way, and the output of the receiver is connected to the Y-plates of a cathode-ray oscilloscope, the X-plates being fed with a d.c. voltage proportional to the frequency deviation at every instant. The complete response curve is thus displayed on the screen of the c.r.t., and the operator can see immediately the effect of any adjustment he makes.

The voltage for the X-plates can be obtained from a circuit of the frequency discriminator type, as used for A.F.C., or may be derived from a simple potentiometer arrangement ganged to the spindle which controls the frequency of the oscillator.

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Exact linearity between this voltage and the frequency deviation is not necessary.

Tuning of Klystron and T.R. Cell

Assuming the receiver is correctly adjusted as regards I.F. tuning and bandwidth, the A.F.C. must be cut out and the manual tuning control, which adjusts the voltage on the repeller electrode of the klystron, set to mid-position.

The tuning plugs on the klystron and T.R. cell are screwed right in, and the transmitter is switched on. The crystal current is adjusted, by varying the coupling to waveguide or coaxial feeder, to some convenient value below the maximum permitted.

The tuning plugs on the klystron are then screwed out until some convenient signal (transmitter breakthrough or a strong fixed echo) reaches a maximum, as shown on the display. The

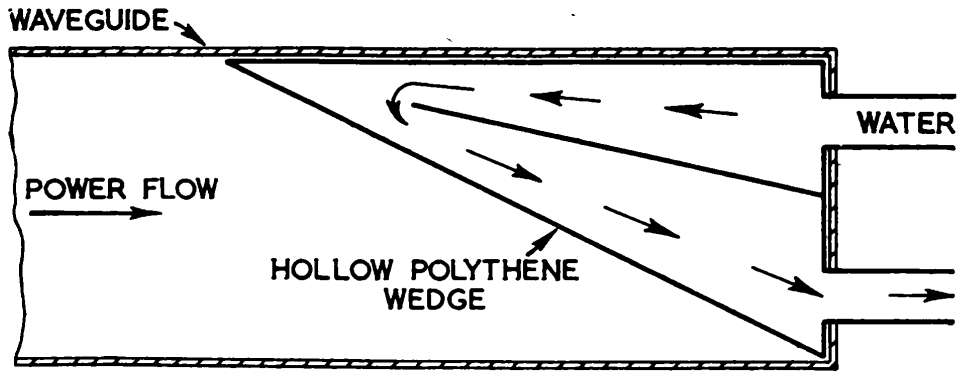

FIG. 17.4.—WATER LOAD FOR WAVEGUIDE.

crystal current usually, but not necessarily, reaches a maximum at the same point. In carrying out this part of the operation, it may be necessary to make some adjustment to the T.R. cell tuning to get sufficient signal.

When the klystron has been tuned as described, the T.R. cell should be similarly dealt with. Finally, the klystron tuning, T.R. cell tuning, and crystal coupling should be varied in turn by small amounts until the best signal to noise ratio on some fixed echo is obtained.

Once the above operation has been correctly carried out, day-to-day adjustment, if necessary, can generally be made by tuning for maximum crystal current. The full procedure should be gone through when a new klystron, magnetron or T.R. cell is fitted.

Equipments not provided with A.F.C. generally have a built-in tuning meter. This may consist of a valve voltmeter circuit connected to the first I.F. stage.

Transmitter Testing

The relative output of a transmitter from day to day can be checked by the echo box as already described. The measurement of absolute power requires more elaborate apparatus.

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One method is to absorb the power in a water load, as in Fig. 17.4 which shows a hollow polythene wedge in the end of a waveguide. The power is calculated from the rate of flow and rise of temperature of the water.

Impedance and Voltage Standing Wave Ratio (V.S.W.R.)

It is difficult to measure impedance at centimetric frequencies, and in practice the impedance of a circuit component is deduced, relative to that of the feeder, by observing the standing wave ratio set up by the circuit element concerned.

Measurement of S.W.R. in a concentric feeder can be made

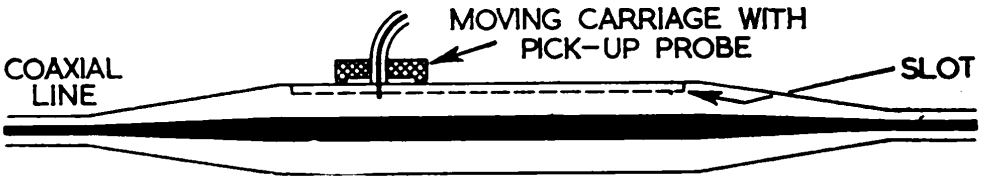

FIG. 17.5.—SLOTTED LINE FOR STANDING WAVE MEASUREMENTS.

by inserting a length of slotted line having a probe carried on a carriage which can be moved along the line over a distance of at least half a wavelength. The voltage picked up by the probe is measured in any convenient manner, e.g. by crystal and microammeter, and the variation of voltage as the probe is moved along the line is noted. The arrangement is shown in Fig. 17.5. A similar device can be used for a waveguide.

For practical convenience the diameters of the conductors in the slotted line are usually greater than in the feeder, but their

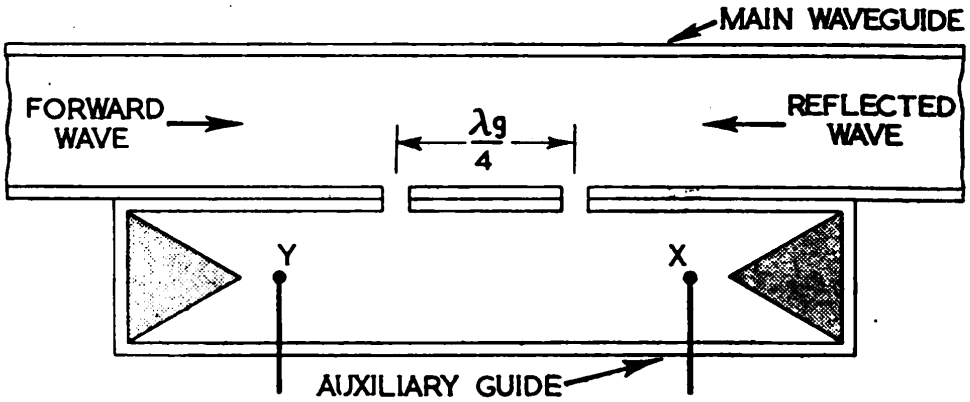

FIG. 17.6.—DIRECTIONAL COUPLER FOR MEASURING FORWARD AND REFLECTED POWERS.

ratio is of the correct value to give the same impedance. The transition from one diameter to the other is tapered to minimize reflection.

From a knowledge of the magnitude of the S.W.R. and the positions of the voltage maxima and minima, the magnitude and phase of the impedance, relative to the feeder, can be calculated from formulae given in Chapter VIII. A complication arises, however, from the fact that, although S.W.R. can be

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measured fairly accurately, it is not always easy to measure phase because the length of the feeder from slotted line to the source of mismatch may not be known with any degree of accuracy. This is not, however, always of importance in practice because the aim is usually to reduce the S.W.R. to unity, when the question of phase no longer arises.

Reflection coefficient may be measured directly by using a short length of auxiliary waveguide connected to the main waveguide by a directional coupler. The relative powers of the forward and reflected waves can then be measured independently—see Fig. 17.6, in which X and Y represent power measuring devices of some kind, e.g. crystals with meters, or thermocouples.

Neon Tube Standing Wave Indicator

This device (Fig. 17.7) can be permanently installed in a transmitter feeder or waveguide and will give an approximate

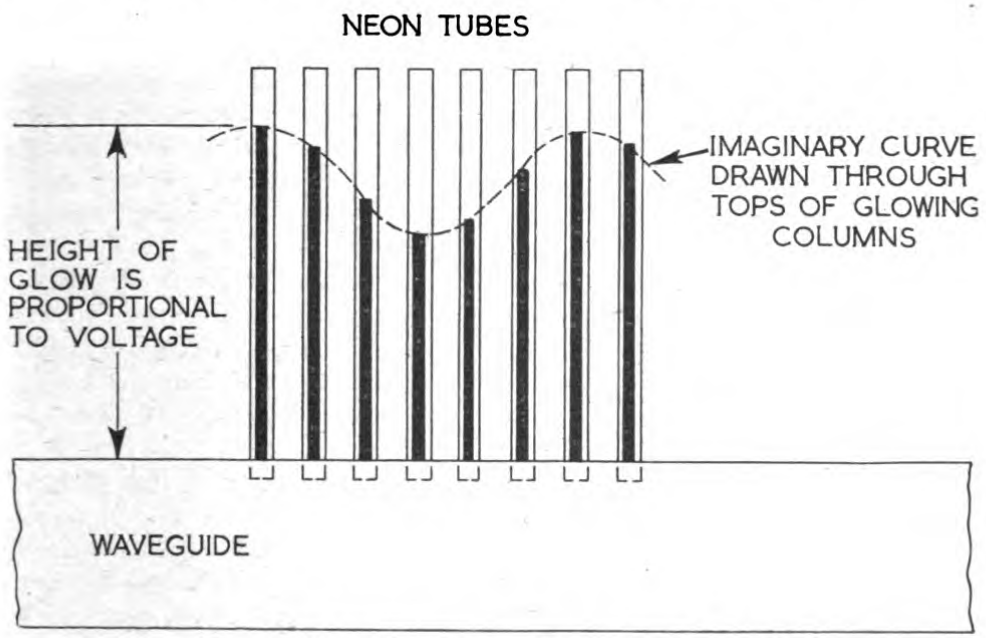

FIG. 17.7.—STANDING WAVE INDICATOR.

visual indication of the variation of voltage along the feeder. It consists of a series of neon tubes spaced at short distances apart over about half a wavelength. The length of the glow in the tube is a measure of the voltage at the point.

Frequency Spectrum of Magnetron

For high grade work it is necessary to ensure that the magnetron is working uniformly on a "single" frequency. This is determined by measuring its spectrum (see Chapter IV, page 22).

A calibrated echo box, whose frequency can be varied over the band concerned, together with crystal detector and indicating meter, may be used.

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APPENDIX

UNITS

THE four main systems of electrical units are :
(*a*) C.G.S. (Electromagnetic), (*b*) C.G.S. (Electrostatic), (*c*) The " Practical " System, and (*d*) The Rationalized M.K.S. System.

(*a*) *C.G.S. Electromagnetic System.* This is based on the centimetre, gramme and second, and on the conception of a " unit magnetic pole ". This is a magnetic pole of such strength that it repels a like pole one centimetre away with a force of one dyne. In the past this system has been used almost universally for scientific purposes, but is now giving way to the M.K.S. system.

(*b*) *The C.G.S. Electrostatic System.* This system is built up in the same way as the C.G.S. electromagnetic system, but starts with a unit charge instead of unit magnetic pole. The system has been used in the past for problems which are mainly of an electrostatic nature, but it is not likely to concern the engineer to any extent.

(*c*) *The Practical System.* This is the well-known system of volts, amperes, ohms, watts, etc. All these units are related to the corresponding units in the C.G.S. (E.M.) system by factors which are powers of 10. The system came into being because some of the C.G.S. units were of inconvenient size for practical use.

(*d*) *The M.K.S. System.* Some fifty years ago, Giorgi pointed out that if the metre, kilogramme and second were taken as the fundamental units of length, mass and time, a system could be built up in the same scientific manner as the C.G.S. system, but with the great practical advantage that the commoner units such as voltage, current, resistance, power, etc., would be identical with the practical units. The electrical starting-point of the system is not a unit magnetic pole, but the ampere, which is defined as that current which, when flowing in two parallel wires one metre apart, produces (*in vacuo*) a force between them of 2×10^{-7} M.K.S. units per metre length. For the purpose of the definition the wires are supposed to be straight, infinitely long, and of negligible cross-section. The M.K.S. unit of force is the newton, and is that force which produces an acceleration of one metre per second per second in a mass of one kilogramme. It is equal to 10^5 dynes (approximately the weight of 102 grammes).

Although the common units in the M.K.S. system are the same as the practical units, it has been necessary to alter certain

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basic definitions. These are:

Permeability (μ) of free space $= 4\pi \times 10^{-7}$

Permittivity or dielectric constant (ϵ) ,, ,, ,, $= \frac{10^7}{4\pi c^2}$

The second of these follows from the first in accordance with the fundamental relation $\mu\epsilon c^2 = 1$, where c is the velocity of light.

The latest value for c is $2{\cdot}9979_2 \times 10^8$ metres per second, but for many practical purposes it is sufficient to take it as 3×10^8.

In the C.G.S. systems, permeability and permittivity of free space are each arbitrarily taken as unity, the former in the electromagnetic and the latter in the electrostatic system.

Unit E.M.F. in the electrostatic system equals $\frac{c}{10^6}$ volts (300, very nearly).

For convenience, permeabilities and permittivities are frequently quoted relative to air as unity. When this is done they should be called relative permeability and relative permittivity, respectively.

RELATIONS BETWEEN M.K.S. AND C.G.S. (ELECTROMAGNETIC) UNITS

	M.K.S.		C.G.S. (e.m.u.)	
	Name	Equivalent	Name	Equivalent
Length . .	Metre	1	Centimetre	100
Mass . . .	Kilogramme	1	Gramme	1,000
Force . . .	Newton	1	Dyne	10^5
Energy (work)	Joule	1	Erg	10^7
Power . . .	Watt	1	—	10^7
Potential (E.M.F.)	Volt	1	—	10^8
Current . .	Ampere	1	—	10^{-1}
Charge . . .	Coulomb	1	—	10^{-1}
Resistance .	Ohm	1	—	10^9
Inductance .	Henry	1	—	10^9
Capacitance .	Farad	1	—	10^{-9}
Magnetic field (H)	Amperes per metre (or ampere-turns per metre)	1	Oersted	$\frac{4\pi}{10^3}$
Magnetic flux (Φ)	Weber	1	Maxwell	10^8
Magnetic flux density (B)	Weber per square metre	1	Gauss	10^4

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One important difference between the practical and the M.K.S. systems lies in the units of flux density. A magnetic flux density of 10,000 lines per square centimetre, in the old system, is identical with unit flux per square metre in the M.K.S.

Examples

(i) An E.M.F. of 1 volt is physically the same as one of 10^8 C.G.S. units.

(ii) A magnetic flux density of one weber per square metre is the same as 10^4 gauss (lines per square centimetre).

Units of Power Measurement

Absolute power is measured in watts, which expression includes kilowatts, megawatts, milliwatts and microwatts.

Relative power or power ratio, e.g. the ratio of the output to the input of an amplifier, feeder system, etc., is measured in decibels.

The ratio of two powers P_1 and P_2 expressed in decibels is $10 \log_{10} (P_2/P_1)$ decibels (db.). This quantity is positive or negative according as P_2 is greater or less than P_1.

The decibel being a logarithmic unit, the overall gain or loss of a system can be found by adding together the gains or losses (measured in decibels) of the constituent parts of the system.

Absolute powers are sometimes expressed in decibels. In such cases the statement means so many decibels above or below some specified power level. For example, 5 dbm. means 5 decibels above one milliwatt.

If voltage or current ratios are expressed in decibels, the formulae are $20 \log_{10} (V_2/V_1)$ or $20 \log_{10} (I_2/I_1)$. These expressions only represent true power ratios if the impedance is the same at the two points concerned.

Another unit, frequently met with in transmission line calculations, is the neper. This is defined as

$\log_e (I_2/I_1)$ or $2{\cdot}3026 \log_{10} (I_2/I_1)$ or $1{\cdot}1513 \log_{10} (P_2/P_1)$ (all these are equal).

Whence 1 neper = 8·686 decibels.
1 decibel = 0·1151 nepers.

FREQUENCY AND WAVELENGTH

(Frequency) × (Wavelength) = (Velocity). If the frequency is in cycles per second and the wavelength in metres, the velocity is in metres per second.

The velocity of radio waves is the same as that of light. In air, it is about one part in 3,000 less than the value given above for free space (*vacuo*).

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Frequency Bands

For convenience, radio frequencies have been divided into bands as follows:

Designation	*Frequency*	*Wavelength*
VLF (very low frequency)	Up to 30 Kc/s	10,000 metres and over
LF (low frequency)	30–300 Kc/s	10,000–1,000 metres
MF (medium frequency)	300–3,000 Kc/s	1,000–100 metres
HF (high frequency)	3–30 Mc/s	100–10 metres
VHF (very high frequency)	30–300 Mc/s	10–1 metres
UHF (ultra high frequency)	300–3,000 Mc/s	100–10 centimetres
SHF (super high frequency)	3,000–30,000 Mc/s	10–1 centimetres
EHF (extremely high frequency)	30,000–300,000 Mc/s	10–1 millimetres

Most modern radars employ frequencies in the UHF and SHF bands, with a few of the older equipments in the VHF band. Intermediate frequencies in radar receivers are generally in the HF or VHF bands.

Some frequency bands have been given alternative designations, those with which radar is mainly concerned being:

L band	390– 1,650 Mc/s	(77–18 cm. approx.)
S band	1,650– 5,200 Mc/s	(18–5·8 cm. approx.)
X band	5,200–11,900 Mc/s	(5·8–2·5 cm. approx.)

MATHEMATICAL NOTES

The base of the napierian or hyperbolic logarithms is $e = 2{\cdot}71828\ldots$

To convert napierian logarithms to common logarithms multiply by 0·4343.

To convert common logarithms to napierian logarithms multiply by 2·3026.

$$e^x = \text{antilog}_{10}\,(0{\cdot}4343x),\quad e^{-x} = \frac{1}{\text{antilog}_{10}\,(0{\cdot}4343x)}$$

(These relations are useful when a table of exponentials is not available.)

$$1 \text{ radian} = 180/\pi \text{ degrees} = 57^\circ\ 17'\ 45''$$

For small angles:

$\sin\theta = \theta$ (radians). Accurate to 1 per cent for angles not greater than 14°.

$\cos\theta = 1$. Accurate to 1 per cent for angles not greater than 8°.

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Trigonometrical Functions (commonly called circular functions)

$$\sin(a \pm b) = \sin a.\cos b \pm \cos a.\sin b$$
$$\cos(a \pm b) = \cos a.\cos b \mp \sin a.\sin b$$
$$\tan(a \pm b) = \frac{\tan a \pm \tan b}{1 \mp \tan a.\tan b}$$
$$\cos^2 a + \sin^2 a = 1 \qquad \sin(\pi/2 \pm a) = \cos a$$
$$\cos(\pi/2 \pm a) = \mp \sin a$$

Hyperbolic Functions

$$\sinh a = \frac{e^a - e^{-a}}{2} \qquad \cosh a = \frac{e^a + e^{-a}}{2} \qquad \tanh a = \frac{\sinh a}{\cosh a}$$
$$\cosh^2 a - \sinh^2 a = 1$$

Complex Functions

$$\sinh(a \pm jb) = \sinh a.\cos b \pm j \cosh a.\sin b$$
$$\cosh(a \pm jb) = \cosh a.\cos b \pm j \sinh a.\sin b$$
$$e^{ja} = \cos a + j \sin a$$

whence

$$e^{jna} = (\cos a + j \sin a)^n = \cos na + j \sin na$$
(de Moivre's Theorem)

PARALLEL PLATE CAPACITOR

A capacitor consisting of two parallel plates, area A square metres, spaced B metres apart, has a capacitance (neglecting edge effects) of

$$\frac{A}{B}\epsilon \text{ farads} \qquad \text{or} \qquad \frac{A}{B}\epsilon\ 10^6 \text{ microfarads}$$

where ϵ is the permittivity (dielectric constant) of the dielectric. For free-space ϵ is $10^7/(4\pi c^2) = 8{\cdot}84 \times 10^{-12}$.

CONCENTRIC OR COAXIAL LINE

a = inner radius of outer conductor.
b = radius of inner conductor.

Capacitance per unit length (1 metre) $= \dfrac{2\pi\epsilon}{\log_e(a/b)}$ farads.

For air dielectric this is very nearly equal to

$$\frac{2{\cdot}41 \times 10^{-5}}{\log_{10}(a/b)} \text{ microfarads.}$$

Inductance per unit length $= \dfrac{\mu}{2\pi}\log_e(a/b)$ henries.

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For air this becomes $4{\cdot}605 \times 10^{-7} \log_{10} (a/b)$.

Characteristic impedance

$$= \frac{1}{2\pi}(\mu/\epsilon)^{\frac{1}{2}} \log_e (a/b) = 2 \times 10^{-7} c \log_e (a/b) \text{ ohms.}$$

For air this is $138 \log_{10} (a/b)$ ohms.

SOME INSULATING MATERIALS USED IN RADAR

Material	*Relative Permittivity (Air = 1)*	*Power Factor*
Bakelite . . .	4–7	0·01–0·1
Distrene (Polystyrene)	2·5–2·6	0·0002–0·0004
Ebonite . . .	2·8	0·006
Glass . . .	5–10	
Mica . . .	2·5–7	0·01–0·07
Mycalex . . .	8·5	0·002
Paxolin . . .	2	0·05
Polythene . .	2·2	0·0006
Pyrex . . .	4·5	0·00017
Tufnol . . .	5	0·03

The above figures are mostly applicable to frequencies of about 1 Mc/s, and should only be regarded as approximate. Different samples may vary somewhat widely.

DECIBEL CONVERSION TABLE

Decibels	Power Ratio	Voltage Ratio	Decibels	Power Ratio	Voltage Ratio
0	1	1	15	31·62	5·623
1	1·259	1·122	16	39·81	6·310
2	1·585	1·259	17	50·12	7·079
3	1·995	1·413	18	63·1	7·943
4	2·512	1·585	19	79·43	8·913
5	3·162	1·778	20	100	10
6	3·981	1·995	30	1,000	31·62
7	5·012	2·239	40	10,000	100
8	6·310	2·512	50	10^5	316·2
9	7·943	2·818	60	10^6	10^3
10	10·0	3·162	70	10^7	$3{\cdot}16 \times 10^3$
11	12·59	3·548	80	10^8	10^4
12	15·85	3·981	90	10^9	$3{\cdot}16 \times 10^4$
13	19·95	4·467	100	10^{10}	10^5
14	25·12	5·012			

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GREEK ALPHABET

α	A	Alpha	ν	N	Nu
β	B	Beta	ξ	Ξ	Xi
γ	Γ	Gamma	o	O	Omicron
δ	Δ	Delta	π	Π	Pi
ϵ	E	Epsilon	ρ	P	Rho
ζ	Z	Zeta	σ	Σ	Sigma
η	H	Eta	τ	T	Tau
θ	Θ	Theta	υ	Υ	Upsilon
ι	I	Iota	ϕ	Φ	Phi
κ	K	Kappa	χ	X	Chi
λ	Λ	Lambda	ψ	Ψ	Psi
μ	M	Mu	ω	Ω	Omega

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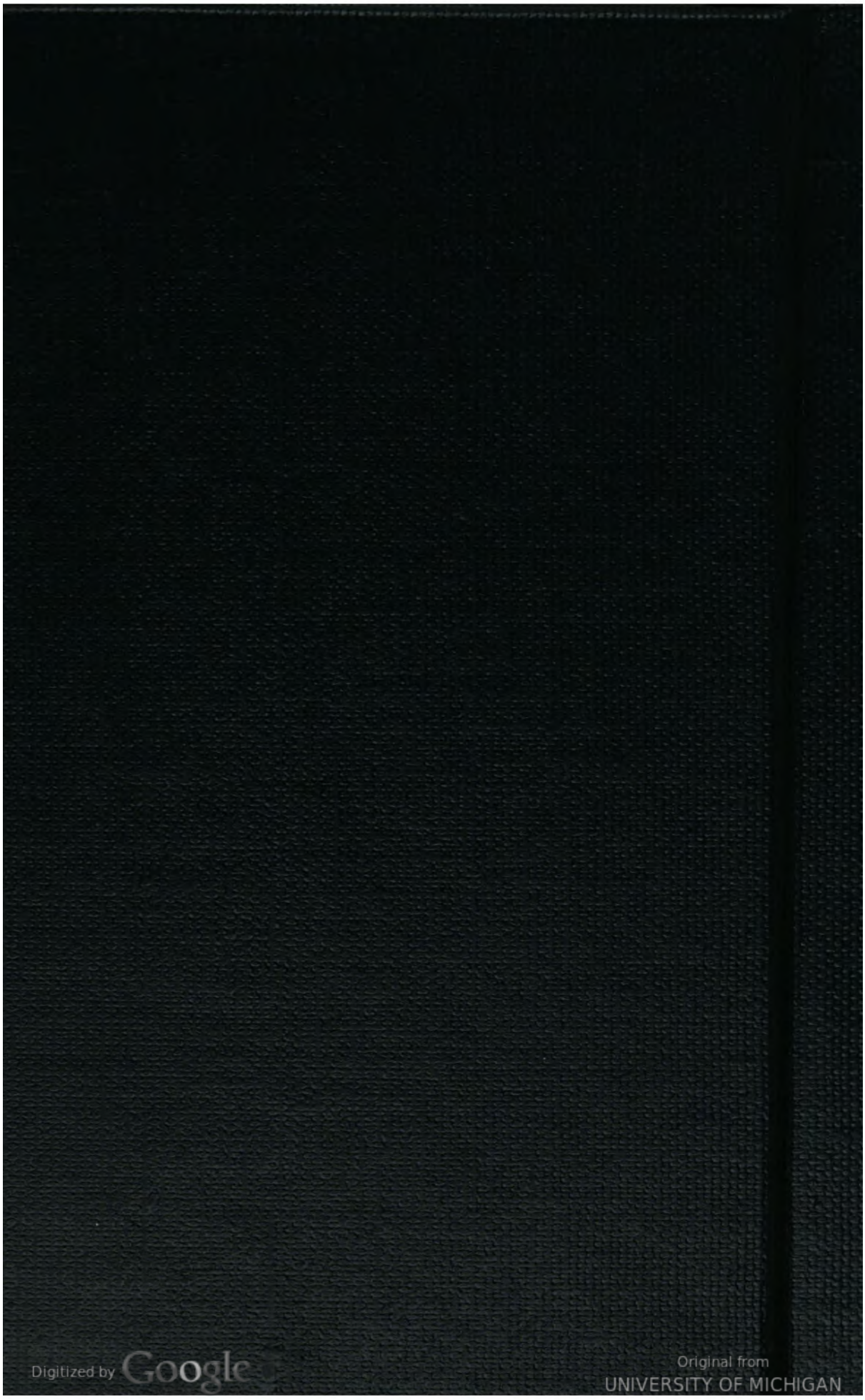
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